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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 17 January, 1929 Number 1

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The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

Subscription rates to the PROCEEDINGS for the current year are received from non-members at the rate of $1.00 per copy or $10.00 per year. To foreign countries the rates are $1.10 per copy or $11.00 per year.

Back issues are available in unbound form for the years 1918, 1920, 1921, 1922, 1923, and 1926 at $9.00 per volume (six issues) or $1.50 per single issue. Single copies for the year 1928 are available at $1.00 per issue. For the years 1913, 1914, 1915, 1916, 1917, 1918, 1924, and 1925 miscellaneous copies (incomplete unbound volumes) can be purchased for $1.50 each; for 1927 at $1.00 each. The Secretary of the Institute should be addressed for a list of these.

Discount of twenty-five per cent on all unbound volumes or copies is allowed to members of the Institute, libraries, booksellers, and subscription agencies.

Bound volumes are available as follows: for the years 1918, 1920, 1921, 1922, 1923, 1925, and 1926 to members of the Institute, libraries, booksellers, and subscription agencies at $8.75 per volume in blue buckram binding and $10.25 in morocco leather binding; to all others the prices are $11.00 and $12.50, respectively. For the year 1928 the bound volume prices are: to members of the Institute, libraries, booksellers and subscription agencies, $9.50 in blue buckram binding and $11.00 in morocco leather binding; to all others, $12.00 and $13.50, respectively. Foreign postage on all bound volumes is one dollar, and on single copies is ten cents.

Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership, etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

Advertising rates to the PROCEEDINGS will be supplied by the Institute's Advertising Department, Room 802, 33 West 39th Street, New York, N. Y.

Changes of address to affect a particular issue must be received at the Institute office not later than the 15th of the month preceding date of issue. That is, a change in mailing address to be effective with the October issue of the PROCEEDINGS must be received by not later than September 15th. Members of the Institute are requested to advise the Secretary of any change in their business connection or title irrespective of change in their mailing address, for the purpose of keeping the Year Book membership catalog up to date.

The right to reprint limited portions or abstracts of the papers, discussions, or editorial notes in the PROCEEDINGS is granted on the express condition that specific reference shall be made to the source of such material. Diagrams and photographs published in the PROCEEDINGS may not be reproduced without making special arrangements with the Institute through the Secretary.

It is understood that the statements and opinions given in the PROCEEDINGS are views of the individual members to whom they are credited, and are not binding on the membership of the Institute as a whole.

Correspondence relative to business, editorial, and advertising matters should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York, N. Y., U. S. A.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered, and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer devisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportions.

Abbreviations—Write a.c. and d.c., kc, μf, μμf, emf, mh, μh, henries, abscissas, antennas Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right, in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

Reprints—With the notification of acceptance of paper for publication reprint order form is sent to the author. Orders for reprints must be forwarded promptly as type is not held after publication.

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RAYMOND A. HEISING
Member Board of Direction, 1929

Raymond A. Heising was born August 10, 1888 at Albert Lea, Minnesota. He received the E.E. degree from the University of North Dakota in 1912, and the M.S. degree from the University of Wisconsin in 1914. Since 1914 he has been a member of the technical staff of the Engineering Department of the Western Electric Company and of its successor, Bell Telephone Laboratories, specializing in radio telephony. He was at the transmitting end of the historical radio telephone experimental transmission between Arlington and Paris, Honolulu and Darien in 1915. During the war he worked on numerous war-time radio projects, and acted as instructor to technical men assigned by the War Department to the Western Electric Company laboratories. He has taken part in practically all of the American Telephone and Telegraph Company's recent radio projects. He has published numerous papers in this and other journals, and holds over fifty patents applying to practical radio development.

The Institute awarded Mr. Heising the Morris Liebmann Memorial Prize for 1921, and elected him a member of the Board of Direction in 1926. He is a Fellow of the Institute, has been Chairman of its Committee on Admissions, and a member of other of its Committees.

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INSTITUTE NEWS AND RADIO NOTES

December Meeting of the Board of Direction

At the meeting of the Board of Direction of the Institute held in the Institute office on December 5, 1928 the following were present: Alfred N. Goldsmith, President; L. E. Whittemore, Vice President; Melville Eastham, Treasurer; John M. Clayton, Secretary; Arthur Batcheller, J. H. Dellinger, R. A. Heising, J. V. L. Hogan, and R. H. Marriott.

The following were transferred or elected to the higher grades of membership in the Institute: transferred to the Fellow grade: H. J. van der Bijl; transferred to the Member grade: P. H. Sohor and Sidney R. Wright; elected to the Member grade: M. W. Kenney, W. T. Runge, and Tadashi Fujimoto.

Ninety-three Associate members and eight Junior members were elected.

Appointment of Assistant Secretary

The Board of Direction of the Institute at its October meeting appointed C. J. Porter, of Buffalo, New York, Assistant Secretary of the Institute. For the past ten years Mr. Porter has been associated with the Westinghouse Electric and Manufacturing Company at Buffalo as radio specialist. He was a member of the organization committee of the Buffalo-Niagara Section and served as Secretary of that Section for two years. Mr. Porter assumed his duties with the Institute on November 1st. He will be mainly occupied in the handling of Section matters.

World Engineering Congress

The Institute has been requested by the American Committee of the World Engineering Congress, which is to be held in Tokio in October of 1929, to sponsor a paper on "The Trend of Radio Broadcasting and its Relation to National Solidarity." A Committee composed of Alfred N. Goldsmith, Chairman; B. A. Clark, B. Ray Cummings, D. G. Little, J. H. Dellinger, E. L. Nelson,

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and Julius Weinberger has undertaken the preparation of a symposium paper which will be read at the Congress and subsequently published in the PROCEEDINGS.

Radio Stations of the World

The November 1928 issue of the PROCEEDINGS contained a list of frequency assignments throughout the world above 1,500 kc. Column three of page 1576 of this list is out of alignment. To give proper readings, move the first twelve lines of column three down one line, and move each line from "Drummondville P.Q." to the bottom of the page up one line.

Reprints of this list can be obtained free of charge by requesting Reprint No. 59. These reprints do *not* contain the above correction.

Chicago Section

The paper on "The Design of Transformers for Audio-Frequency Amplifiers with Preassigned Characteristics," by Glenn Koehler, published on page 1742 of the Decenber, 1928 issue of the PROCEEDINGS was presented before the April 20, 1928 meeting of the Chicago Section.

1929 Annual Convention

Plans for the 1929 Annual Convention to be held in Washington, D. C. on May 13–15 are being completed. Chairmen of the various committees have been appointed as follows: F. P. Guthrie, Convention Chairman; Thomas McL. Davis, Registration and Arrangements; Professor A. E. Kennelly, Fellowship; S. S. Kirby, Trips; F. P. Guthrie, Banquet; R. D. Heinl, Publicity.

Special arrangements are being made to entertain lady guests at this Convention. A Committee composed of Mrs. F. P. Guthrie, Mrs. J. H. Dellinger, Miss Mary T. Loomis, and Miss Elizabeth M. Zandonini has been appointed to arrange a number of features of special interest to the ladies present during the Convention.

The technical program plans call for a symposium meeting on the technical problems involved in radio legislation, and a similar meeting for a symposium on still and moving picture transmission by radio.

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As the annual meeting of the U.R.S.I. consisting of a full day of technical programs follows on the 16th, the entertainment features of the Institute convention are being stressed. Members desiring to remain in Washington to attend the U.R.S.I. meeting will be cordially welcomed. At the U.R.S.I. meetings a large number of excellent radio papers are always presented.

Among the trips tentatively scheduled for the convention are inspection trips to the Naval Research Laboratory, the Bureau of Standards, Arlington Radio Station and the Tomb of the Unknown Soldier, and Mount Vernon.

The completed program of the convention will be announced in an early forthcoming issue of the PROCEEDINGS. Members are urged to set aside May 13 to May 15 for the 1929 Convention!

Dewey Decimal Classification

With this issue, on the title page of each paper will be found the Dewey Classification of each paper published in the PROCEEDINGS. This number appears as a footnote.

The index to the PROCEEDINGS for 1927 and 1928 will be published in the 1929 Year Book, arranged according to the Dewey Decimal Classification.

Members desiring further information as to the use of this method of classification and indexing are referred to "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, which may be obtained from the Superintendent of Documents, Government Printing Office, Washington, D. C. for ten cents.

Edison Medal Awarded to Frank B. Jewett

The Edison Medal has been awarded by the Edison Medal Committee of the American Institute of Electrical Engineers to Dr. Frank B. Jewett, Vice President of the American Telephone and Telegraph Company, and President of the Bell Telephone Laboratories, for his contributions to the art of electrical communication.

The Edison Medal was founded by associates and friends of Mr. Thomas A. Edison, and is awarded annually.

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U. S. Government Call Book

The June 30, 1928 edition of "Commercial and Government Radio Stations of the United States" is now available from the Superintendent of Documents, Government Printing Office, Washington, D. C. for fifteen cents.

In addition to a list of the land and ship radio stations, both commercial and governmental, the 1928 edition contains two lists of broadcasting stations, the first being the assignments prior to November 11, 1928 and the second the assignments subsequent to that date. The broadcast station lists are arranged alphabetically by call letters, alphabetically by states and cities, and numerically by wavelengths and frequencies.

Standard Frequency Transmissions by the Bureau of Standards

The Bureau of Standards announces its schedule of radio signals of standard frequencies for use by the public in calibrating frequency standards and transmitting and receiving apparatus. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington, which goes into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to about 500 to 1,000 miles from the transmitting station.

The transmissions are by continuous wave radiotelegraphy. The signals have a slight modulation of high pitch which aids in their identification. A complete frequency transmission includes a "general call" and "standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues for about two minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about four minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 171, which

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may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

RADIO SIGNAL TRANSMISSIONS OF STANDARD FREQUENCY SCHEDULE OF FREQUENCIES IN KILOCYCLES

Eastern Standard Time	Jan. 21	Feb. 20	March 20
10:00–10:08 P.M.	125	500	1500
10:12–10:20	150	600	1700
10:24–10:32	200	650	2250
10:36–10:44	250	800	2750
10:48–10:56	300	1000	2850
11:00–11:08	375	1200	3200
11:12–11:20	450	1400	3500
11:24–11:32	550	1500	4000

Institute Meetings

BOSTON SECTION

The Boston Section held a meeting in Cruft Laboratory, Harvard University, Cambridge, Massachusetts on November 16th. Dr. G. W. Pierce, Chairman of the Section, presided.

A paper, "Electrical Transmission of Pictures and Images," was presented by J. W. Horton, Chief Engineer of the General Radio Company.

Following the presentation of the paper the following members participated in its discussion: A. F. Murray, H. W. Lamson, A. E. Kennelly, H. J. Tyzzer, G. W. Pierce, L. F. Curtis and P. B. Bauer.

On December 12th a meeting of the Boston Section was held in Cruft Laboratory to hear a paper by Major Richard H. Ranger, of the Radio Corporation of America, on "Recent Developments in Photoradio."

It is hoped that both of these papers may appear in early issues of the PROCEEDINGS.

CANADIAN SECTION

A meeting of the Canadian Section was held on October 10th in the Electrical Building, University of Toronto. A. M. Patience, Chairman of the Section, presided.

B. F. de Bayley presented a paper, "A Direct Reading Audio-Frequency Bridge." Following the presentation the following members of the section participated in its discussion: Messrs. Patience, Smith, Meredith, Thompson, and others.

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Fifty-three members of the Section and guests attended the meeting.

Cleveland Section

The second meeting of the 1928–29 season was held on November 23rd, in the auditorium of the Physics Building, Case School of Applied Science.

Chairman John R. Martin presided.

Dr. Dayton C. Miller, Ambrose Swasey, Professor of Physics at Case School of Applied Science, presented a paper "The Physical Characteristics of Music and Speech." The paper explained the early experiences of the author, which led him into the study of sound. Dr. Miller explained the development of the "phonodeik" to make visible the character of sound and waves. It consists of a glass diaphragm linked with a minute pivoted mirror. A beam of light is reflected by the movement of the diaphragm and then spread out on a screen by a revolving mirror.

A resumé of the underlying principles of sound and music was given, and the speaker explained and demonstrated tonal quality and wave composition by means of ten tuning forks.

The talk was supplemented with lantern slides, which were explained in detail. The "phonodeik" was demonstrated and projected a moving image of sound picture on a large screen in front of the audience. The wave shape produced by various musical instruments and the human voice was clearly shown.

Following the presentation of the paper a large number of persons participated in the discussion. There were about one hundred and thirty-five members of the Cleveland Section present and questions were asked at the end of the meeting.

New York Meeting

Dr. A. S. Eve, of McGill University, Montreal, Canada, came to New York on the evening of December 5th to present the paper "Reception Experiments in Mount Royal Tunnel," by Messrs. Eve, Steel, Olive, McEwan, and Thompson.

The paper appears in an early forthcoming issue of the Proceedings.

Following its presentation the following members took part in the discussion: A. S. Eve, Alfred N. Goldsmith, R. H. Marriott, Haraden Pratt, J. H. Dellinger, G. W. Kenrick, and others.

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One hundred and seventy-five members and guests attended the meeting.

The January 2nd, 1929 meeting in New York will be addressed by Dr. V. Zworykin, of the Westinghouse Electric and Manufacturing Company, on the subject of "Facsimile Picture Transmission."

PHILADELPHIA SECTION

On November 23rd a meeting of the Philadelphia Section, held in the Franklin Institute, was addressed by G. W. Kenrick. The paper was entitled, "The Heaviside-Kennelly Layer and Its Relation to Radio Transmission Phenomena."

The paper reviewed the historical development of radio transmission theory and considered evidence for the existence of a conducting layer. Group time and phase retardation measurements in determining the virtual height of the layer as a function frequency and angle were described. New group time experiments which show the diurnal variation of the layer and the effect of a magnetic storm were presented. The paper also described the methods of determining the ratio of "real" to "virtual" height of the layer from group time and phase retardation experiments, and the application of the results thus determined to a study of the variation of electronic and ionic density in the upper atmosphere as a function of height was mentioned.

The meeting was presided over by J. C. Van Horn, Chairman of the Section. Thirty-five members of the Section were present.

PITTSBURGH SECTION

A meeting of the Pittsburgh Section was held on November 20th in the Fort Pitt Hotel, Pittsburgh. L. A. Terven, Vice-Chairman of the Section, presided.

V. D. Landon, of the Westinghouse Electric and Manufacturing Company, presented a paper on "Radio Installation for Apartment House Use." The paper described an antenna distribution system on radio frequencies for apartment houses and similar structures. The system is composed of a radio-frequency amplifier and a distributing coupling circuit. A single antenna approximately 30 feet high and 80 feet long is used with the system in conjunction with a separate frequency amplifier in each group of ten or less receivers. Each receiver

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has its own distribution coupler. Detailed theory of the system was explained and worked up from formulas.

Following the presentation of the paper, Messrs. Sutherlin, Sunnegrin, McKinley, Allen, and Terven participated in its discussion.

Twenty-four members of the Section attended the meeting.

Washington Section

On November 8th a meeting of the Washington Section was held in the Continental Hotel, 1 Capitol Street, Washington, D.C. F. P. Guthrie, Chairman of the Section, presided.

Warren B. Burgess, of the Naval Research Laboratory, presented a paper on "Radio Compass in Theory and Practice."

Following the presentation of the paper, the following took part in its discussion: H. G. Dorsey, G. D. Robinson, D. G. Howard, Colonel Parrott, F. P. Guthrie, T. Parkinson, R. B. Stewart.

Forty-six members and guests attended the informal dinner preceding the meeting and sixty-five members attended the meeting.

On December 13th a meeting of the Washington Section will be held in the Continental Hotel. Dr. E. O. Hulburt, of the Naval Research Laboratory, will deliver a paper on "Radio Transmission and Magnetic Storms."

Committee Work

Committee on Sections

A meeting of the Committee on Sections was held in the Institute Office on November 16th. The following were present: Donald McNicol, Chairman; Arthur Batcheller, and C. J. Porter, Assistant Secretary.

The Committee studied the proposed revised constitution, which is to be submitted to the Institute membership for approval in the near future, with a view to making necessary recommendations to the Board of Direction regarding portions of the Constitution affecting the Institute Sections.

Correspondence looking to the formation of Institute Sections in Schenectady, St. Louis, Milwaukee, Minneapolis, and Cincinnati was reviewed.

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Committee on Admissions

The Committee on Admissions met on December 5, 1928 in the Institute office. R. A. Heising, Chairman, E. R. Shute, and F. H. Kroger were present. The Committee considered six applications for transfer or election to higher grades of membership in the Institute.

Personal Mention

L. E. Hayslett has left Kenmore, New York, to become Inspection Engineer of the United Reproducers Corporation of Rochester, New York.

A. C. Matthews has left Schenectady, New York for Chicago where he is now employed as radio engineer with the Stewart Warner Corporation.

W. E. Miles has been transferred from the Naval Communication office, Navy Yard, Boston to the Naval Radio Traffic Station, Bar Harbor, Maine.

R. V. Beshgetoor, formerly with the General Electric Company at Schenectady is now in the Engineering Products Division of the Radio Corporation of America in New York City.

Edgar H. Felix has joined the staff of the National Electrical Manufacturers' Association, specializing in radio problems. Mr. Felix is on the Institute Committee on Meetings and Papers.

W. W. Lindsay, Secretary of the Los Angeles Section of the Institute, is now associated with the Movietone Department of William Fox Studios at Hollywood, California, as research engineer.

William F. Diehl, who for the past ten years has been connected with the A. H. Grebe Company as Chief Engineer, recently joined the staff of the Radio Division of the Victor Talking Machine Company, of Camden, New Jersey.

Maurice Berger, formerly with the Radio Corporation of America at New York City is now associated with the Electrical Research Laboratories, Incorporated, of New York City. Mr. Berger has been a member of the Institute's Committees on Membership and Sections for several years.

David Casem, for the past five years radio editor of the New York Telegram has been appointed director of the publicity bureau of broadcasting station WOR at Newark. Mr. Casem has served on the Institute's Commitee of publicity for a number of years.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED DECEMBER 5, 1928

Transferred to the Member grade

California	Los Angeles, 800 North Spring Street	Sohor, P. H.
England	Shipley, Yorkshire, 14 Bankfield Drive	Wright, Sidney R.

Elected to the Member grade

Illinois	Elmhurst, 265 W. North Avenue	Kenney, M. W.
Germany	Berlin, Wittenau, Robertstr	Runge, Wilhelm T.
Japan	Tokyo, Setagaya, 429 Taishido	Fujimoto, Tadashi

Elected to the Associate grade

Arkansas	Blytheville, 304 Davis Avenue	Lintzenich, C. L.
California	Glendale, 408 Wing Street	Shomler, H. B.
	Hollywood, 1030 N. Western Avenue	Finstein, Paul
	Hollywood, 1345 North New Hampshire Avenue	Sarver, Frank M.
	Los Angeles, 4503 Paulhan	Andersen, H. W.
	Los Angeles, 1444 So. Norton Avenue	Buffinger, William
	Los Angeles, 1441 W. 51 Place	Hill, Alfred LeRoy
	Los Angeles, 1050½ Cloverdale	Stevens, Sterling M.
	Sterling City	Overton, D. C.
	Stockton, 1520 N. Eldorado Street	Holt, Pliny E.
Georgia	Atlanta, c/o General Electric Co., P.O. Box 1698	Bussey, H. E.
	Atlanta, 1331 Lucile Avenue, S. W.	Wallace, Lawrence T.
	Columbus, c/o Columbus Electric and Power Co.	Farris, Herbert Talley
Illinois	Chicago, 100 W. Monroe St.	Fillmore, Harry Philip
	Chicago, 2247 Calumet Avenue	Hauser, Albert
	Chicago, c/o R. C. A., 100 W. Monroe Street	Kennedy, W. J. B.
	Chicago, 1544 North Dearborn	Nourse, Philip R.
	Chicago, 4430 Clifton Avenue	Turner, C. G.
	Dixon, 521 McKenney Street	Hall, Howard J.
	Kewanee, 219½ West Second Street	Larson, Clifford L.
	Urbana, Dept. of Elec. Eng., University of Illinois	Hershey, Arthur W.
Indiana	Indianapolis, 1036 Eugene Street	Byers, Russell R.
	Valparaiso, 402 Monroe Street	Alexander, Leslie Alpheus
	Valparaiso, 555 So. College Avenue	Imler, Glenn M.
Iowa	Hale, Box 29	Kruse, Gerald D.
Louisiana	Baton Rouge, 830 North Street	Webre, Andrew S.
	New Orleans, 3624 Magazine Street	de Otter, G.
	New Orleans, 1222 Eagle Street	Pasquet, Jean E.
	New Orleans, 2655 Canal Street	Rabito, A. Paul
Massachusetts	Cambridge, Massachusetts Institute of Technology	Barrow, W. L.
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	Cambridge, General Radio Co.	Worthen, Charles E.
	North Attleboro, Old Post Road	Barrett, Kerman R.
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	Swampscott, 12 Humphrey Terrace	Humphrey, Stanley M.
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	Buffalo, 176 Laurel Street	Wright, George A.
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	New York City, 72 Cortlandt Street, Room 301	Moore, E. B.
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	Ravenna, Box No. 4	Rettig, Floyd A.
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	Mobile, P. O. Box 801	Helt, Scott
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	Los Angeles, 444 Kings Road	Grosjean, G. M.
	San Francisco, 2370 Filbert St.	Smith, Herbert H.
	San Jose, 398 N. 8th St.	Kelliher, John Edward
Colorado	Colorado Springs, 920 E. Monument	Harvey, Francis M.
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	Cleveland, 2845 Prospect Ave.	Umbrecht, Stanton
	Lakewood, 1288 Ramona Ave.	Drummond, P.
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	Newphiladelphia, 334 Minnich Ave.	Nolan, John
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Ohio	Lakewood, 12974 Emerson Ave.	Thompson, J. Kent
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Part II

TECHNICAL PAPERS

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

A DIRECT READING RADIO-FREQUENCY METER*

By
R. C. Hitchcock
(Research Laboratory, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Penna.)

Summary—*The method used is the well known one of reading a radio frequency by measuring the beat note produced with a calibrated crystal oscillator.*

A new type of direct reading audio-frequency meter having a scale of 2.0 to 4.5 kc per second makes the device automatic.

The meter scale divisions are 0.1 kc apart, and under reasonable conditions, the accuracy of the reading is of this order.

THERE are two classes of meters which are usually used to measure radio frequencies. The first is the tuned circuit absorption type. With this type the radio source must be coupled closely enough to supply the necessary energy to the meter. The first wavemeters of this type used a single variable condenser shunted by a single coil and some kind of resonance indicating device. Well made tuned circuit meters are reliable to about 2.5 kc per second in the present broadcast band.

The Bureau of Standards Frequency Meter Type B was the next step in the absorption type of meter, the range being narrowed with a corresponding increase in precision by shunting a small variable condenser across a large main fixed condenser. Two series coils were used, the smaller one being unwound during calibration, so that the wave to be measured came at the mid point of the small variable condenser.

Recently a still more precise narrow range frequency meter has been placed on the market.[1] This meter has been designed to cover 0.3 per cent of a radio station's frequency with the use of a single control. At 1500 kc per second this means that the whole range of the meter is 4.5 kc per second. The manufacturer specifies 10 scale divisions per kc, and guarantees for six months a precision of 500 cycles per second when the instrument is kept within ±5 deg. F of the calibration temperature given on the chart. If a more accurate meter of this type is desired, the whole unit might be calibrated in a temperature-controlled oil bath. Stand-

* Dewey decimal classification: R210. Original manuscript received by the Institute, October 16, 1928.

[1] Frequency meter, type 532, General Radio Co.

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ard oscillator crystals are usually temperature-controlled, and a standard tuned circuit would be more accurate if temperature control were employed. An ingenious resonance indicating device makes the instrument much more sensitive than previous meters of the absorption type.

Oscillator as a Frequency Meter

The second type of frequency meter is a calibrated oscillator. The beat note between the oscillator and the desired source is measured and applied to the known value of the oscillator frequency. This type of meter has the advantage that it can be operated some distance from the radio source to be measured.

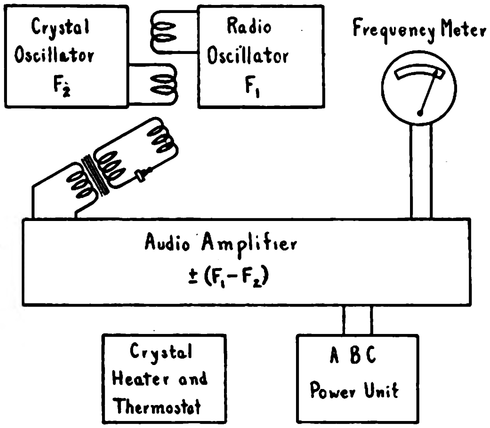

Fig. 1—Schematic Diagram.

By using suitable audio amplification, several miles may separate the radio source and the standard measuring oscillator. It is the purpose of this paper to describe a meter of this second type, which can give a radio frequency directly on a meter scale.

It is realized that the general scheme of using a beat method of calibrating one frequency from another is a standard one, and that harmonics of a standard oscillator are also widely used for calibration purposes. However, as the use of an indicating audio-frequency meter has received little attention, the circuit and various details will be given here.

Standard Crystal Oscillator

This meter being a beat frequency indicating device, it is essential that an accurately calibrated crystal oscillator unit be

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employed. Such units have been described[2] and will not be discussed here. In the schematic drawing of Fig. 1 the crystal oscillator unit generates the radio frequency F_2. The source of radio frequency to be measured F_1, is loosely coupled to F_2 and the beat frequency

$$f = \pm(F_1 - F_2) \qquad (1)$$

is passed on to the audio amplifier from a detector circuit which is tightly coupled to the crystal oscillator. This beat frequency

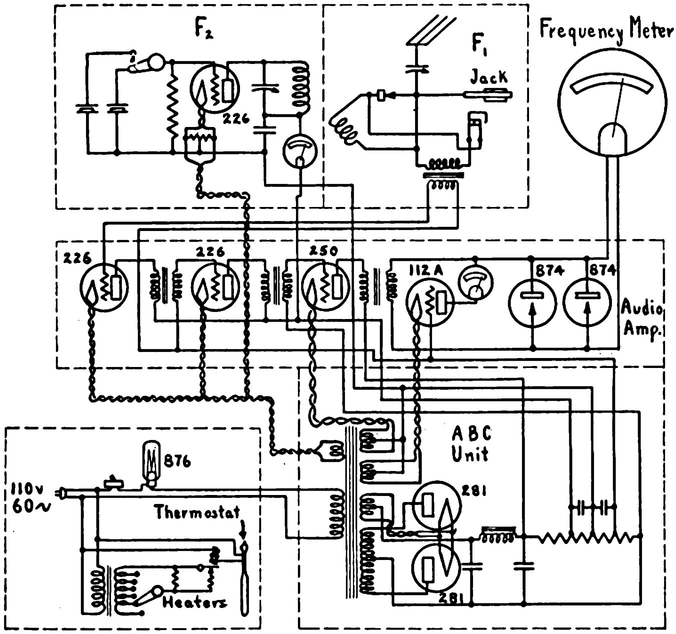

Fig. 2—Wiring Diagram.

after being amplified is fed to the indicating audio-frequency meter. This meter can be calibrated either in terms of the audio frequency of the beat (Fig. 3), in which case the meter reading is applied to the calibrated crystal frequency to determine the frequency of the source; or the scale can be calibrated directly in radio frequency (Fig. 4), having the same kilocycle range as the audio frequency.

The special audio-frequency meter which made this device

[2] Crossley, Proc. I.R.E., **15**, 9; January, 1927. Worrall and Owens, Proc. I.R.E., **16**, 778; June, 1928.

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possible[3] has a useful scale range of 2.0 to 4.5 kc per second spread over an arc of 90 deg. This meter is of the tuned circuit type, its impedance being three times as great at the center of the scale as at the ends. To minimize this effect on the rest of the audio circuit, a 2:1 stepdown transformer is placed between the UX-250 tube and the meter. The complete wiring diagram is given in Fig. 2. The voltage amplifiers are UX-226 tubes, and the crystal oscillator also uses a UX-226 tube. Across the frequency meter are placed two UX-874 voltage regulator tubes to improve regulation, as will be mentioned later.

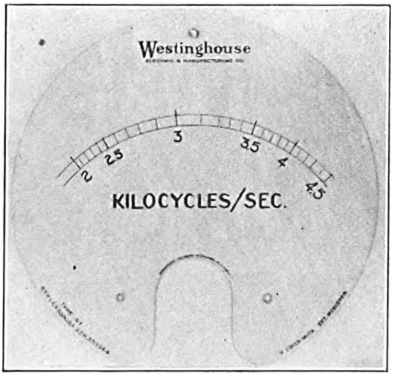

Fig. 3

Voltage Regulation at the Meter

The reading of the audio-frequency meter is dependent to some degree on the voltage supplied. Fig. 6 gives the characteristic curve of the meter when the applied voltage is kept at 60 volts. Where the slope is least, in the middle of the scale, the readings will be most accurate. The arbitrary scale plotted as the abscissa is a uniform one which was spaced along the arc of the meter scale for experimental purposes. Due to the low impedance at the ends of the scale it was not possible, without building another special amplifier, to maintain the voltage above 60 throughout the whole range. Hence this curve of Fig. 6 at 60 volts was taken as a basis of comparison to show the effects of changing voltage. Fig. 7 gives the deviation in cycles per second from the values of Fig. 6 for 70, 80, and 90 volts applied to the

[3] Designed by Mr. B. E. Lenehan, Supply Eng., W. E. and M. Co., Newark, N. J.

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meter. It will be noted that the maximum deviation is less than 200 cycles per second with voltages from 60 to 80. The use of two UX-874 tubes in parallel across the meter was found to keep the voltage to 68 ± 5 volts, therefore the variations due to voltage, over the range 2.0 kc to 4.0 kc are of the order of ±50

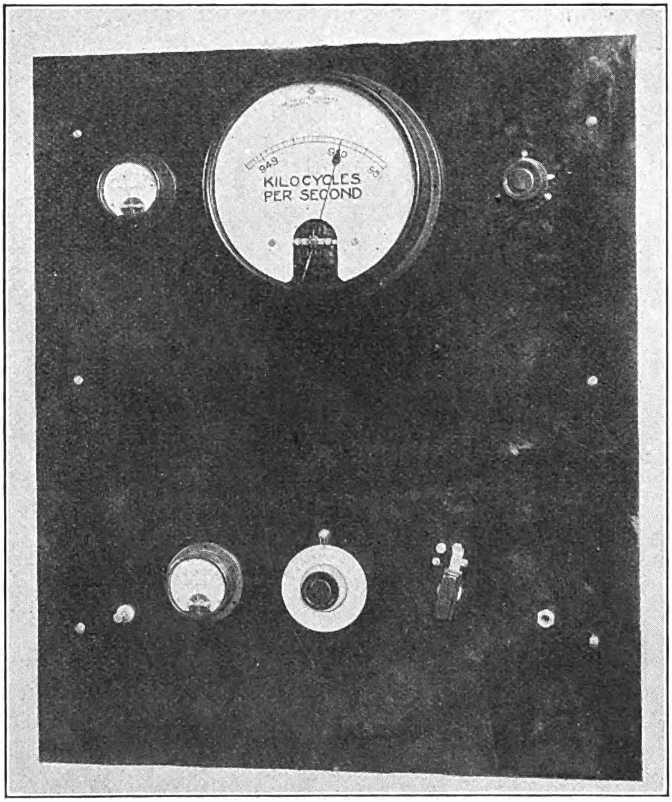

Fig. 4

cycles per second.[4] Closer regulation of voltage is hardly justifiable in view of the other factors affecting the frequency; the crystal temperature, oscillator tuning and variations in coupling to the source. A more nearly constant voltage for precision work could be obtained by using a manually controlled resistor across the meter. Such a resistor was the first scheme used to keep the

[4] The meter indication changes slightly with room temperature variations.

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applied meter voltage nearly constant. The resistor knob is shown in the upper right of Fig. 4. The alternating voltage across the meter can be read by a calibrated milliammeter in the plate circuit of a UX-112A tube. The connection is given in Fig. 2. After inserting the two UX-874 regulator tubes, the manual

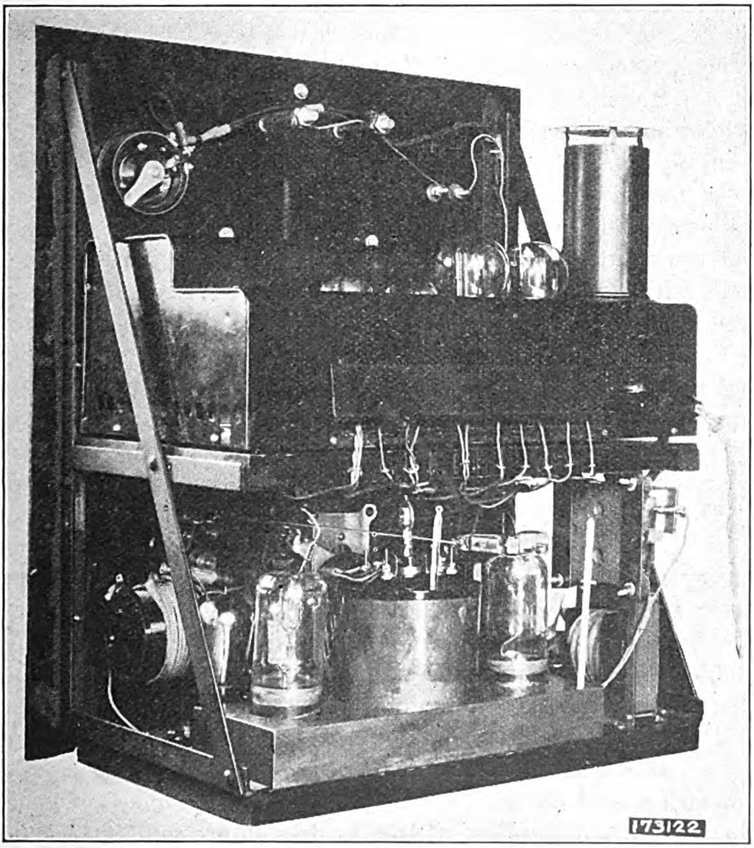

Fig. 5

resistor was seldom used except for fine adjustments. For this reason it was not included in Fig. 2.

It is not necessary that the fundamental frequencies of the standard and the source to be measured should be within an audio frequency of each other, as beats may be obtained between harmonics. For instance, a 50 kc per second standard crystal can be used to read station frequencies close to $50n$ kc per second

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where n is an integer between 11 and 30 for stations in the broadcast band of 550 to 1500 kc per second. To do this, f, the beat frequency, should be within the range of the frequency meter, the relation between F_1, f, and the 50 kc standard being given by (2)

$$f = \pm(F_1 - 50n) \qquad (2)$$

The proper harmonic of the standard will beat with the station frequency and the reading will appear on the meter scale, which can be translated directly to a radio-frequency value. If a high harmonic is used, sufficient amplification should be provided to supply the meter with the required voltage.

In measuring a properly designed crystal-controlled radio-frequency source with this frequency meter it is quite unlikely that the change should ever be more than a fraction of a kilocycle, which would be indicated by the meter. But if a tuned circuit master oscillator is being measured it is just possible that its frequency might change by such an amount that the audio beat would be the same as when the radio frequency was at its correct value. In this case, the frequency meter would indicate the proper beat, but on the other side of the standard crystal oscillator.

To illustrate, suppose the frequency of the checking crystal to be 947 kc in order to check a radio station at 950 kc, the frequency assigned to KDKA in June 15, 1927. The frequency meter would also read this beat, 3 kc, if the station frequency were 944 kc, which would be 6 kc from the assigned value. This double value is indicated by the plus or minus signs of (1) and (2). To eliminate this ambiguity two crystal standards can be used; to check a 950 kc station, crystals of 947 and 953 kc per second would serve. Then for the case cited, the second crystal standard would not give its check at its regular place of 3 kc, when the low frequency of 944 kc was impressed, but would tend to show 9 kc, and although this value is off scale for the frequency meter, the operator can tell by the action of the milliammeter that either a low or a high frequency is being measured. That is, the impedance of the meter being low at both ends, the low milliammeter reading would mean a low impressed voltage on the meter, and in turn would indicate a frequency not on the scale of the meter. Another way of eliminating the ambiguity of the double sign would be to use two crystal standards on the same side of the measured frequency, for instance 947 kc and

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948 kc. Then for a 950 kc station the correct reading would be 3 kc for the first, and 2 kc for the second. Should the source change its frequency to 944 kc, the first standard would indicate its usual 3 kc, but the second would now show 4 kc definitely locating the value of radio frequency at 944 kc. The two crystals used in the meter are shown in Fig. 6 mounted in the copper-glass tubes described by the author in an earlier paper.[5] A large copper block serves to conduct heat from the thermostatically controlled chamber in the center. The wiring details are incorporated in Fig. 2 at the lower left. The main heater is left on all the time, while the auxiliary one is turned on and off by a

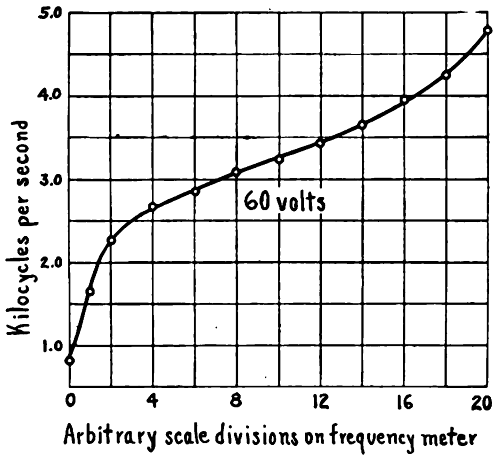

Fig. 6

contact making thermometer and suitable 110-volt relay. The switch lever and two points for selecting either of the two standard crystal oscillators are shown in Fig. 4 at the lower right.

When the meter is being used to read a radio station frequency while a program is being broadcast, the crystal detector supplies the program, as well as the beat frequency, to the audio amplifier. If the program consists of a man speaking at a frequency of a few hundred cycles per second and the beat is, say 4 kc per second, the frequency meter will momentarily take up a position between these two values. The higher frequency of a woman's voice will not change the reading so much. However, there are always

[5] R. C. Hitchcock, Proc. I.R.E., 15, 902; November, 1927.

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momentary pauses in a program when the beat frequency will be the predominant one, and the meter can always be read with a precision better than one scale division.

If only a high-frequency beat is to be used, a high pass filter has been used to cut out the lower voice frequencies satisfactorily. This limits the use of the meter to its high end, but reduces the variations of the pointer when indicating the beat.

In the original setup of this frequency meter, KDKA is measured whenever it is on the air, a relay being automatically operated by the energy collected by a tuned antenna circuit. The meter as ordinarily balanced has an equilibrium position at about the center of the scale. This was not desired when the

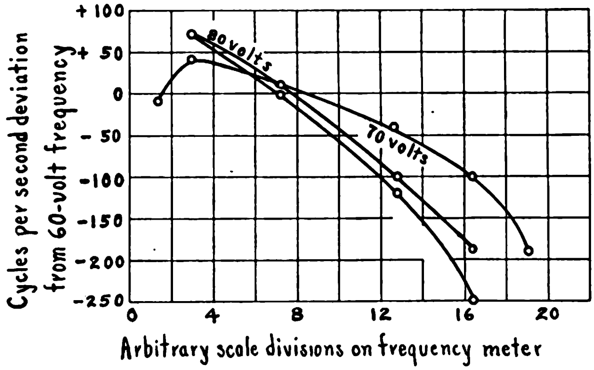

Fig. 7

scale was calibrated in radio frequency, as it intimated that the station was on the air at all times, and indicated a fictitious frequency whenever the station was not being operated. To make the meter pointer move to the extreme left, a back contact was added to the relay operating the amplifier, so that when the station went off the air, the relay opened the power circuit to the amplifier and connected the 110-volt 60-cycle supply through a 2 μf condenser across the frequency meter. This drew the meter pointer off scale to the left whenever the station was not on the air.

To make the meter available for other purposes than station frequency checking, a jack has been placed as shown in the wiring scheme of Fig. 2 and photograph of Fig. 4 so that audio energy within the range of the meter may have its frequency measured, and the crystal oscillator automatically disconnected by inserting a plug.

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Warning Devices for Frequency Changes

As this meter is probably accurate to better than 200 cycles per second under adverse conditions, and as the Federal Radio Commission specifies a maximum deviation of 500 cycles per second from an assigned broadcast frequency, it is possible to devise warning instruments which will operate when the pointer of the frequency meter touches light contact members placed

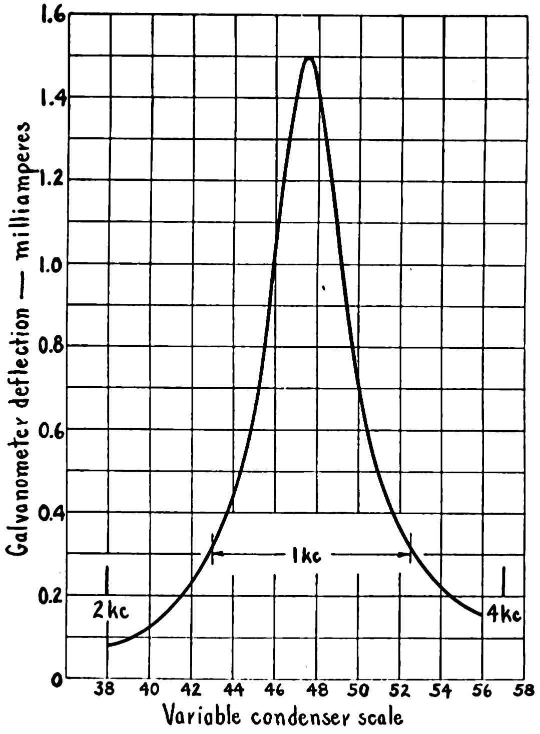

Fig. 8

within the allowable limits of frequency deviation. A grid glow relay[6] which operates with a fraction of a microampere would serve to operate heavier current relays to sound suitable alarms when the frequency changed by a predetermined amount.

Parallel Tuned Circuit

Another indicating device which could replace the audio-frequency meter would be a parallel resonant circuit. This circuit would be tuned to the beat frequency *f*, and a meter

[6] D. D. Knowles, *Elec. Jour.*, 176, April, 1928.

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showing the change in oscillating current used to read frequency changes. A typical curve with a peak at 3 kc is shown in Fig. 8. A 6-mh coil and variable condenser of 0.05 μf across a fixed one of 0.5 μf was used. A loosely coupled pick-up coil, detector, and d.c. milliammeter gives the resonance indication without adding to the resistance of the resonant circuit.

By using resistance coupled amplification any beat frequency could be measured by this scheme, although accuracy is greater for lower beat frequencies. A tuned circuit meter when used to measure an audio beat is more accurate than another tuned circuit meter with the same sharpness of resonance,[7] at radio frequencies, by the ratio of the radio to the audio beat frequency. This assumes the use of a good crystal oscillator standard for producing the beat frequency.

An advantage of this latter scheme over the frequency meter is that the standard could be chosen several kilocycles away from the frequency to be measured, so that the operator could be sure that the frequency was either higher or lower than that of the standard oscillator. This would permit the measuring of several radio frequencies with a single crystal standard by plugging in suitable tuned audio circuits. This also means that the precision would be reduced, as the beat frequency was increased.

Conclusion

This frequency meter is automatic, requiring no manipulation of controls to read a radio frequency. Its scale is spread out so that the frequency divisions are 0.1 kc apart. By using several standard crystals, the meter can read several frequencies and their harmonics, with a precision depending in part on the calibration of the standard crystal. A final advantage is that the meter need not be placed as close to the source as a tuned circuit meter.

Of course the meter described is more complex than a tuned circuit meter, but it is felt that a meter of this type would be of service to radio stations and laboratories. Simple relay schemes can be devised to turn on the meter whenever the radio frequency is on, requiring a minimum of attention.

The writer wishes to express his thanks to Mr. V. E. Trouant for his helpful suggestions and for his data on the parallel tuned circuit; to Mr. E. B. Landon for his cooperation at KDKA; and to Mr. J. C. Batchelor for much of the experimental work connected with this frequency meter.

[7] Circular of the Bureau of Standards, No. 74, Ed. 1918, pp. 36–37.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

ON THE DETERMINATION OF THE OPTIMUM RADIATION ANGLE FOR HORIZONTAL ANTENNAS*

By

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Summary—*A standard of (directional) radiation at 15 meters was constructed consisting of a vertical half-wave antenna placed one-half wave above the earth, suitably excited, placed in front of a reflector consisting of five reflector wires slightly more than one-half wave long on a parabolic surface of opening of one wavelength. This was found to give directional characteristics showing a broad maximum of field strength twice that of the simple antenna without reflector.*

A parabolic reflector and radiating system rotatable about a horizontal axis perpendicular to the line to the distant station were then constructed near Berlin. The antenna at the focal axis of the paraboloid was one wave long, one-quarter wavelength from the parabola apex; and the parabolic reflector was of one wavelength opening and consisted of 9 wires along the paraboloid each approximately a half-wave long.

The entire system was rotated while reception audibility measurements were made at Buenos Aires, a 2-kw transmitter being employed. The horizontally aimed system was found most effective. This held for both 15 and 20 meters. Similar results were obtained for reception on directional antennas. The conclusion is drawn that horizontal short-wave radiation is most desirable.

IN our first experiments with horizontal antennas in August and September, 1925,[1] it was found necessary for the best technical utilization of the effect of horizontal antennas to concentrate the antenna radiation at the angle most advantageous to reception. At first we tried to obtain this by combining several horizontal antennas in a horizontal plane (the antennas were $\lambda/2$ apart) with the corresponding surface of reflectors $\lambda/4$ under them. The various antennas had to be excited with a definite phase difference between them for adjustment of a certain angle of emission. We had no difficulty in setting the different phases among the antennas, but we had no phase indicator to show just what phase was actually present in each antenna. We were, therefore, compelled to return to the simple horizontal antenna system and to concentrate radiation by means of a parabolic reflector. The first parabolic reflector was made of sheet copper for a wavelength of 11 meters. When turn-

* Dewey decimal classification: R125.6. Original manuscript received by the Institute, June 29, 1928. Translation received July 19, 1928.

[1] A. Meissner, "On Space Radiation," *Jahrbuch fuer drahtlose Telegraphie*, **30**, p. 77; *Telefunken-Zeitung*, October, 1927.

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ing the reflector for this wavelength, we found the optimum angle at 38 deg. and a second, somewhat less favorable angle at 80 deg. The reflector could only be depressed to an angle of about 35 deg. below the horizontal. We, therefore, set up a fixed reflector system, which could be remodeled easily for radiations of 0 deg., 10 deg., 20 deg., and 30 deg., for purposes of comparison with the rotatable reflector. The comparison reflector was a wire reflector. Tests had in the meantime shown that the sheet could be replaced by wires; the wire reflector was even somewhat superior to one made of sheet. At the test wavelength of 11 meters and at angles of radiation of 0 deg., 10 deg., and 20 deg., the fixed reflector did not show any definitely noticeable difference as compared with the metal reflector at 38 deg.

The ensuing tests were to determine the optimum radiation angle for operating wavelengths; wavelengths above 15 meters.

Tests with the 20-meter Wave

The existing reflector was then enlarged for the longer wavelengths. At each end another arch was added at a distance of 5 meters to the three parabolic wooden arches supporting the reflector surfaces, so that an antenna two half-waves long and oscillating in phase could be built in for a wavelength of 20 meters. The antenna was hung in the focal line of the reflector (5 meters from the apex of the parabola—width of the parabola opening: 20 meters). The parabolic surface was formed by 9 reflector wires (9.60 meters long). Preliminary experiments had determined the optimum length of the reflector wires as well as the concentration of radiation obtainable with the reflector system employed. This was done with a fixed vertical parabolic reflector system with an opening 15 meters wide and a focal length of about 3.75 meters which corresponded to the horizontal rotatable reflector. The tests were made at a wavelength of 15 meters. The transmitting antenna was vertical, $\lambda/2$ long, with its center $\lambda/2$ above the earth excited at the lower voltage loop by a double line system $4/\lambda$ long wires 15 cm apart. One of the energy supply lines ended $\lambda/4$ over the ground, while the antenna, $\lambda/2$ long, was directly connected to the other line. In the investigation of the influence of the number and tuning of the reflector wires it was found that for a parabola with an opening of λ width, five reflector wires suffice and that an increase of the number of wires does not improve concentration. When

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changing the length of the reflector wires one gets a broad maximum of field intensity in front of the reflector. The optimum value lies at a natural wavelength of the reflector wires approximately 5 per cent longer than half the operating wavelength. The tests were made at a distance of about 10 wavelengths with the aid of a tuned tube receiver. In Fig. 1 the numerical values are proportional to the field intensity. We see that the concentration is comparatively broad. The field intensity in front of the reflector is practically constant throughout an angle of approxi-

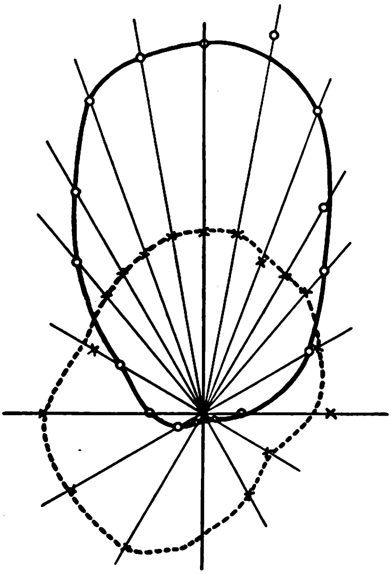

Fig. 1—Radiation Diagram for the Optimum Adjustment of the Reflector.
————— with reflector
- - - - - - without reflector

mately 40 deg. At the maximum it is about twice as large as the field intensity of the vertical antenna alone, represented by the broken line. The dissymmetry of this curve is probably due to the nearness of the antenna system. Within six minutes the reflector could be turned from 40 deg. above the horizontal in the direction of Buenos Aires through 90 deg. to 40 deg. above the horizontal in the other direction, away from the receiver. The transmitter was a self-oscillating push-pull vacuum-tube transmitter fed with 500-cycle alternating current. The antenna was connected through a double line (wires 12 cm apart) 30

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meters long. Standing waves in the line were avoided by means of a suitable coupling of the energy supply line to the antenna. The power in the antenna was about 2 kw. Buenos Aires measurements were made by the shunted-telephone audibility meter method. The angle tests were made on two days (Nov. 30 and Dec. 1, 1927) between 11 A.M. and 12 noon and between 6 P.M. and 9 P.M. by means of the continuous turning of the reflector within the above-mentioned angles in the space of six minutes. A total of 38 observation values was obtained. The mean of the reception results corresponds approximately to the curves of

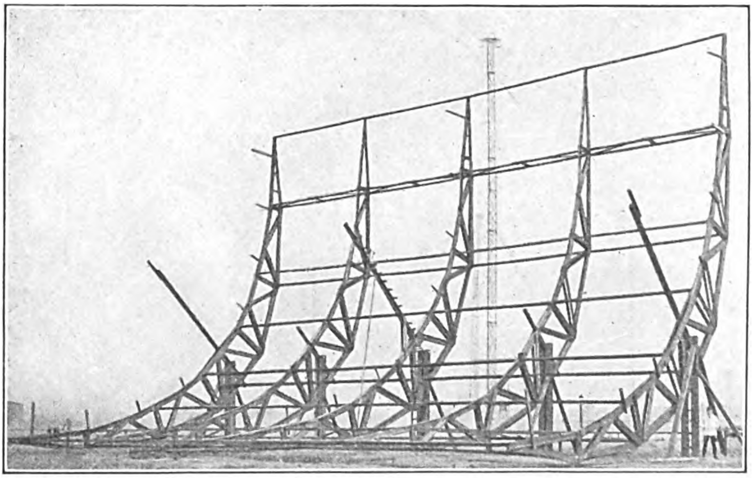

Fig. 2—Rotatable Reflector in the Setting for A Radiation Angle of 40 Degrees.

Fig. 4. The unbroken line represents the tests on the first day, the broken line those on the second day. The reception results on both days show that the optimum radiation of the rotatable reflector takes place at its smallest angle: 40 deg.–50 deg. In order to see whether even smaller radiation angles than 40 deg. are still better, we again arranged a fixed reflector system alongside the rotatable one and exactly similar to it. Within an hour it could be easily set at any angle from 0 deg. to 40 deg. The fixed reflector was always compared with the optimum setting of the rotatable reflector (40 deg.). The first thing we noticed was that for the same radiation angle the fixed reflector was always superior to the movable one, about in the ratio 1.3 to 1. This was perhaps due to the fact that there were losses in the wooden

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parts of the rotatable reflector system. Probably it is also due in part to the circumstance that the antenna of the fixed system was somewhat higher ($h=11$ meters) than that of the rotatable reflector ($h=8$ meters).[2] The table gives the values for the fixed and rotatable reflectors at various settings of the fixed reflector:

TABLE I

Rotatable reflector	40 deg.	1
Fixed reflector	40	1.3
" "	30	1.5
" "	20	2.0
" "	10	2.5
" "	0	5.0

This shows that the radiation angles between 0 deg. and 20 deg. are superior to the radiation angles of the rotatable reflector between 40 deg. and 50 deg. The fact that the angle 0 deg. is so particularly favorable is possibly due to the fact that the very dispersed radiation of the reflector (see Fig. 1) is concentrated more strongly by the reflecting influence of the earth. In general, in the emission of horizontally polarized waves the influence of the earth on the upward bending of the rays is apt to be very great.[3]

Tests with the 15-meter Wave

The reception results with the rotatable reflector at $\lambda=15$ meters gave the same results as at the 20-meter wave. The rotation tests were made on December 16th and 17th, 1927, between 12 noon and 3 P.M., Central European time. The plotted results obtained by turning the reflector were altogether like those with the 20-meter wave (Fig. 4). The reception maximum lay at 40 deg. Here again the rotatable reflector was compared with the fixed reflector at the settings 0 deg., 10 deg., and 20 deg. The reception values in this comparison were not as definite as with the 20-meter wave, because the method of audibility measurement at the receiving end had been changed and was unreliable.

[2] With a width of the parabola opening equal to the wavelength, the fixed horizontal antenna with parabolic reflector could not be hung lower than 11 meters, on the other hand the antenna of the rotatable reflector could not be arranged higher than 8 meters without considerably increasing the cost of the system. Comparative tests of reception between a simple horizontal antenna hung at a height of 8 m and a similar antenna hung at a height of 11 m gave differences smaller than 1:2. When used in combination with the reflector, the difference may be estimated to be smaller than 1:1.5.

[3] As a result of tests of simple vertical dipoles hanging at various heights T. L. Eckersley had assumed as early as 1926 that for vertical antennas the horizontal radiation of the transmitter was most advantageous. (See *Jour.* I.E.E., 65, 601, 1927.)

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But the observed values were clear enough to show that reception at horizontal emission was at least twice as good as at a radiation angle of 40 deg. These conclusions were confirmed by tests with the regular operating transmitter AGA (wavelength 15 meters); by changing the angle from 40 deg. to 0 deg. the received signal intensity increased from 3 to 4 times. These results have also been verified by tests made during the erection of the new Telefunken antennas.

The knowledge thus gained—that the optimum radiation of horizontal antennas is most advantageous when radiation is approximately parallel to the earth's surface—proved that we

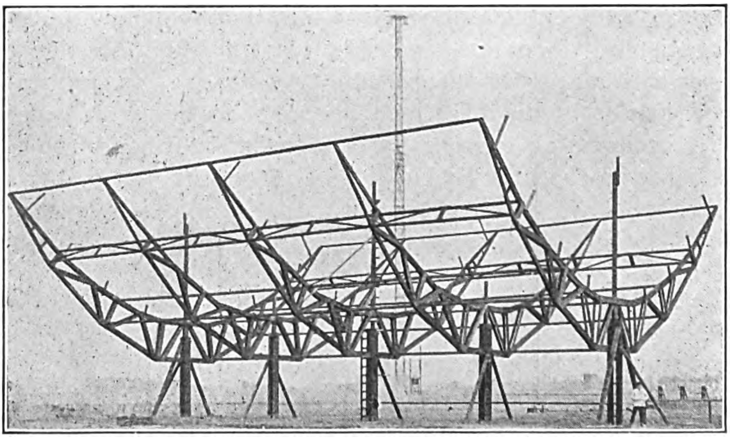

Fig. 3—Rotatable Reflector in the Setting for A Radiation Angle of 90 Degrees.

could dispense with the parabolic reflector, as horizontal radiation can be achieved more easily with plane antennas and reflectors. Tests with such antennas were made parallel to those described above. We shall report on these tests later on.

We also endeavored to determine the optimum angle of incidence of the horizontal component of the received wave by means of the above mentioned rotatable reflector, shown in Figs. 2 and 3, together with the fixed reflector system built alongside it for horizontal radiation. A modern short-wave receiver (two stages of r.f. and two stages of a.f. amplification with a separate heterodyning oscillator) was connected to the energy line of the antenna hanging in the parabola. The tests were carried out by Herr A. Gothe. During reception the reflector was

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turned from 40 deg. to 120 deg. When turning it was found that reception remained unchanged in the region from 40 deg. to about 60 deg. From 60 deg. to 90 deg. and 120 deg., signal intensity continuously diminished. Reception with the fixed reflector set at 0 deg. was just noticeably better than the reflector at 40 deg. (The antenna of the fixed reflector hung at a height of 11 meters while that of the rotatable reflector hung at a height of 8 meters.) The ratio of signal intensity for the reception angles 0–40 deg.

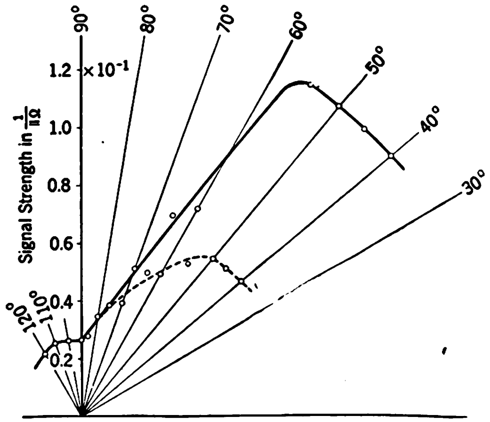

Fig. 4—Signal Strength in Reciprocal Parallel Ohms. Tests on Nov. 30 and Dec. 1, 1927.

to that for the angle 90 deg. is about 4 to 1. We can thus assume that the incoming horizontal radiation affects the antenna system at a small angle, i.e., we can use the same antenna arrangements for the reception of the horizontal component of the incoming wave that were found most suitable for the transmitter. Of course the above reception test is only of limited value. The tests could be made only during a few hours in the forenoon while the large transmitter in Nauen was idle since the rotatable reflector is set up on the grounds of the Nauen station.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

MAGNETOSTRICTION OSCILLATORS*

BY
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(Rumford Professor of Physics, Harvard University, Cambridge, Mass.)

Summary—*The present paper describes a newly discovered method of using magnetostriction to produce and control electrical and mechanical frequencies of oscillations in a range of frequencies extending from a few hundred cycles per second to more than three hundred thousand cycles per second. The method involves the interaction of the mechanical vibrations of a magnetostrictive rod and the electric oscillations of an electric circuit in such a way that the electric currents in the circuit stimulate the rod to longitudinal vibration by magnetostriction and the vibrations of the rod react by magnetostriction on the electric circuit to maintain constancy of frequency.*

The constancy of frequency obtained compares favorably with that obtained with the piezo-electric crystal oscillators. For ease of construction and operation the magnetostriction oscillator has a great advantage over the piezo-electric oscillator, particularly in that the construction and adjustment of the magnetostriction vibrators is so simple that large numbers of standards of frequencies all operable with the same electrical circuits may be had at small expense.

The magnetostriction oscillators supply a particularly pressing need in the range of frequencies below twenty-five thousand cycles per second in which range crystal control is impractical on account of the expense of obtaining sufficiently large crystal vibrators. In the range between twenty-five thousand cycles per second and three hundred thousand cycles per second the magnetostriction oscillators and the crystal oscillators have a common field of usefulness. At frequencies greater than three hundred thousand cycles per second, the magnetostriction oscillators (although active up to two million cycles per second) are feeble with the present arrangement of apparatus. Their harmonics may, however, be employed up to frequencies of several millions per second.

This account contains also methods of calibration of the vibrators and their use in the calibration of wavemeters and frequency meters, data on the velocity of sound in various metallic alloys, data on the elastic constants of metals, including their temperature coefficients, description of methods of sound production, and a theoretical investigation of sound propagation in a viscous magnetostrictive medium.

Circuits and Mounting of Magnetostriction Oscillator for Controlling Oscillations in a Vacuum-Tube Circuit. As shown in Fig. 1, two coils are employed, one, L_1, in the plate circuit

* Dewey decimal classification: R210. Abstract presented at the meeting of the Boston Section, Institute of Radio Engineers, March 16, 1928; original paper presented to the American Academy of Arts and Sciences, Boston, January 11, 1928; reprinted from *Proc.* Amer. Acad., **63, 1, 1928 by special arrangement with the Academy.**

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and the other, L_2, in the grid circuit of a high-μ vacuum tube. The magnetostriction rod, R, which we may call the vibrator, is placed axially within the coils, and rests centrally on a support between the coils, or if desired is held in a clamp at the support. In the diagram the right-hand coil is in the plate circuit in series with a B-battery. The left-hand coil is in the grid circuit. A variable air condenser C is connected between the plate and grid, so as to be across both coils. Sometimes the condenser C is connected across only one of the coils.

By means of plugs and sockets the double-coil unit may be replaced by others of different inductances. With a given coil

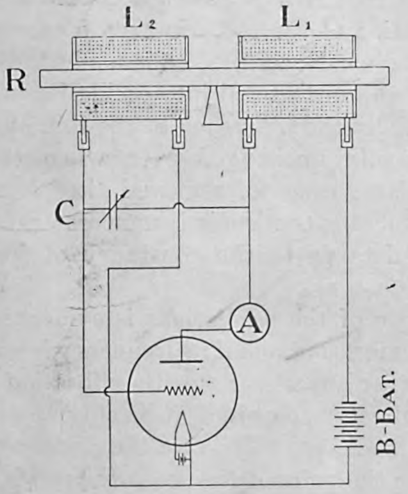

Fig. 1—Oscillator Circuit.

unit many different rods may be used by merely pulling out one and putting in another. The rod is free within the coils, which may have a clearance as large as the diameter of the rod itself. The rod is magnetized permanently by the plate current, or by a permanent magnet placed near it.

A direct-current milliammeter at A serves to indicate the plate current, and by its change with change of C to indicate the presence of oscillations.

Reversed Coil Note. One fact is to be noted that for the best operation of the apparatus as a constant-frequency device, the coils are so wound and oriented that a steady current flowing from the filament to the plate and a steady current, if there were

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one, flowing from the filament to the grid produce magnetic fields in the same direction along the axis of the coils. This sense of the winding of the coils with respect to each other is opposite to that of the familiar electric oscillator circuits; but it should be noted that such an arrangement as that here shown, with certain values of circuit constants, will oscillate electrically without the rod or with the rod restrained.

Operation of the Magnetostrictive Rod in Rendering the System Oscillatory. A system such as is shown in Fig. 1 may be entirely non-oscillatory when the rod is restrained from vibration by being held or when the condenser C has a value far removed from the value required to give the circuit a period near the period of vibration of the rod. In such a case when the rod is released and the condenser has, or is made to have, the proper value, the rod and circuit fall into oscillation with a frequency which is essentially the frequency of the rod, and the frequency remains practically unchanged even when the condenser is varied over a large range or removed altogether. With proper choice of the coils the condenser is unnecessary.

Numerical data as to the constancy of frequency will be given below.

The existence of the oscillations is evidenced by the sound emitted by the vibrating rod, if its frequency is within the audible range; or, whether audible or not, the vibration is evidenced by the change of direct-current indication of the plate milliammeter. When the rod is allowed to vibrate the plate current changes to double or triple the value it has when the rod is restrained.

Operation of the Magnetostrictive Rod in Stabilizing the Frequency of an Electrically Oscillatory System. If we consider the circuit of Fig. 1, and especially if we note that the electrical feed-back between grid coil and plate coil is the reverse of that usually employed in producing electrical oscillations, we may understand that with a given choice of condenser, coil-winding, and coil-spacing, the system may or may not be electrically oscillatory when the magnetostrictive rod is restrained from vibration.

In the preceding paragraph note is made of the action with a non-oscillatory electrical system, which is the preferable mode of operation of the magnetostrictive system when employed with low-frequency vibrations of say five hundred to three thousand cycles per second.

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At higher frequencies ranging from three thousand to three hundred thousand cycles per second, it is more convenient, and just as reliable, to allow the system to be electrically oscillatory even when the rod is restrained, and to employ the magneto-

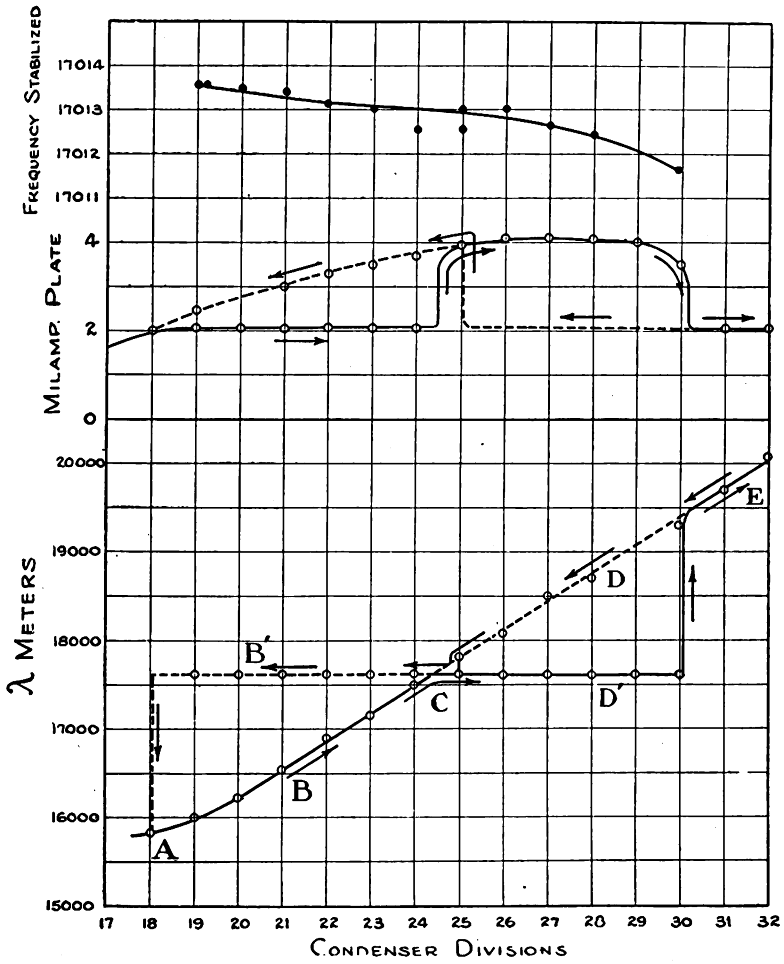

Fig. 2—Illustrating Stabilization of Frequency.

strictive rod merely to stabilize an already existing frequency of the electrical system when the latter is independently adjusted to a value near resonance with the period of mechanical vibration of the rod. This is done as follows: Any given rod, say one of frequency of twenty thousand, is placed in the coils. The

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plate milliammeter is observed as the condenser C is varied. At a certain value of C, the plate current suddenly jumps to a large value (from one and one-half to three times its previous value). This indicates the incidence of rod-vibrations. The condenser may now be adjusted through a considerable range with no material change of the electrical or mechanical frequency of the system. It may be left oscillatory for days without changing frequency although the electrical characteristics may undergo large changes in that time.

General Note on Materials of the Magnetostrictive Rods. This research, which has now extended over some years, has concerned itself largely with the study of materials for the vibrators. The qualities required are large magnetostrictive effects and constancy of mechanical frequency in spite of changes of temperature and of intensity of magnetization, coupled with constancy of frequency in spite of changes of condenser settings, vacuum-tube characteristics, plate currents, and filament currents.

For these purposes pure iron and irons with various carbon contents are relatively useless, as having too small a magnetostrictive effect.

Pure nickel on the other hand is a good vibrator with, however, some lack of stabilizing power in that detuning slightly affects the frequency.

Alloys of nickel and iron in certain proportions are good vibrators, especially those having about 36 per cent nickel and 64 per cent iron (that is, of about the constitution of invar and Stoic metal). This alloy has, however, a large temperature coefficient of frequency.

Alloys of chromium, nickel, and iron are good vibrators, and for many purposes commerical Nichrome is one of the best materials easily available.

An alloy of nickel and copper known as Monel metal, containing 68 per cent Ni, 28 per cent Cu, and small percentages of Fe, Si, Mn, and C, is a very powerful oscillator. It usually has too small a residual magnetism to oscillate without an auxiliary polarizing means.

Alloys of cobalt and iron are strong vibrators.

Tubes of nickel make good oscillators for qualitative work and demonstration purposes and for sources of sound, but are somewhat lacking in constancy as frequency-stabilizers. Such

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tubes of nickel wholly or partly filled with lead or type metal permit the easy construction of low-frequency vibrators because the velocity of sound in lead is small and gives a low-frequency of longitudinal vibration without excessive lengths of the rods.

By using a tube of material that has a negative temperature coefficient of frequency (such as nickel), in combination with a tight-fitting internal core of metal that has a positive temperature coefficient of frequency, (such as stoic metal), I have made composite vibrators of frequency practically independent of temperature. Other composite vibrators are described below.

All of the vibrators are made more powerful by annealing.

Mechanical Vibrations Longitudinal. In most of the applications of the magnetostrictive vibrator to the production and stabilization of frequency I have employed the longitudinal vibrations of the rod either in its fundamental mode or some harmonic mode of longitudinal vibration. In the designation of constants in this paper, unless otherwise stated, reference is to the fundamental mode of longitudinal vibration. It is usually not worth while to use harmonic modes of vibration of the rods since new rods of any desired frequency can be made and calibrated with such ease that the harmonic mechanical vibrations are interesting only in studies of the mechanics of the rod itself. Investigations of this character are now under way.

By sending a current lengthwise through the rod so as to give a circumferential magnetic field and at the same time applying a current to the solenoid, it is theoretically possible to produce torsional oscillations in the rod and these may be used to stabilize frequencies. I have found evidence of these torsional vibrations, but because of the complication of apparatus involved, these torsional oscillations are at present left out of the account.

Vibrators Magnetically Polarized. The rods are polarized by a steady state of magnetization upon which the oscillatory magnetizing forces are superposed, so that the resulting state is an increase of magnetization when the oscillatory current is in one direction and a decrease of magnetization when the oscillatory current is in the other direction. These two states are accompanied by an increase (or decrease) of length and a decrease (or increase) of length, respectively. If the rod were not thus polarized the change of length would be the same for each

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half cycle of the current and would result in a mechanical vibration of double periodicity.

The polarization of the rods is usually attained by permanently magnetizing them in advance by a solenoid traversed by a strong direct current. With most substances the remanent magnetism is sufficiently strong to keep the rods sensitive as oscillators through years of normal use as standards of frequency in spite of any demagnetizing effects of the oscillations. The plate current of the tubes also assists in the polarization provided the rods, if permanently magnetized, are always inserted with their poles in the proper direction, as may be indicated on the coil mounting.

In cases of metals, like Monel metal, which have very small residual magnetism, I find that a small horse-shoe permanent magnetic placed in the vicinity of the rod is sufficient to give the required polarization, even when the small horse-shoe permanent magnet is 15 cm from the rod. With large oscillatory currents the polarization should be correspondingly large so that reversals of magnetization do not occur.

First Step in Explaining Action of Magnetostrictive Rod. Along with the description of the apparatus occasional paragraphs are introduced explanatory of the theory of operation, with a final mathematical treatment of the system. Magnetostriction is the distortion of a body (in this case the lengthening or shortening of the rod) when magnetized. A rod of nickel when magnetized shortens by about one one-millionth of its length for a magnetizing field of one gauss. This is the observed fact when the magnetization is produced by a steady magnetizing force. This shortening is thus very small, for the reason that the shortening must take place against the enormous elastic force of the body. On the other hand, when the specimen is magnetized by a force that increases and decreases in an oscillatory manner at a period resonant with the period of the body the shortening and lengthening may be more than one hundred times as great as that obtained with the same magnitude of constant current applied to the specimen. In the resonant oscillatory case the contraction and expansion are no longer opposed by the elastic force of the body but only by its viscosity. We may call the decrease or increase of length under magnetization the *direct effect.*

There is also an *inverse effect.* When a magnetized bar is stretched mechanically its state of magnetization is changed.

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We may now apply these facts to the system of Fig. 1. Any small fortuitous change of current through the plate coil L_1 changes the magnetization of the rod and causes it to be deformed (lengthened or shortened). This deformation is propagated along the rod to its left-hand end and exists temporarily as a deformation within the coil L_2. The deformation changes the state of magnetization and consequently induces an electromotive force in L_2. This acts on the grid and produces an amplified current change in the plate circuit and in L_1. The oscillating currents in the system thus build up to a large amplitude with a frequency determined by the frequency of longitudinal mechanical vibration of the rod.

Sample Exhibit of Rods. Fig. 14 is from a photograph of several rod vibrators. *G* is a set of Stoic metal vibrators with a range of frequencies extending from ten thousand to thirty thousand cycles per second at intervals of one thousand cycles per second. These are described in Table I. *H* is a set of rods of Nichrome for each thousand cycles extending in range from twenty-six thousand to fifty-seven thousand cycles (see Table II). *F* is a rod of Nichrome with a frequency of 117,705 cycles per second. *D* shows a nickel tube filled with type metal. On account of the low velocity of sound in the type metal the specimen has a frequency of two thousand cycles per second, while a solid nickel rod or a nickel tube of the same length would have a frequency of nearly four thousand. *E* is a part of a similar tube, which is 94.35 cm long, and is filled with lead. It has a frequency of one thousand cycles per second, which can be adjusted over a range of some seven cycles by clamp weights that may be moved nearer together or farther apart to get a higher or lower frequency. By using these movable weights this rod or a harmonic of it can be brought to beat zero with other standards. *C* is a solid oscillator of stainless steel (about 11 per cent chromium). It has a frequency of 23,640 cycles per second, and is here shown to call attention to the fact that rods of large diameters do not fail to vibrate even at high frequencies. In fact the specimen *A* is a thin disc between corks in a glass tube and while 1.5 cm in diameter is only about 2 mm long and has a period of about a million and a half cycles per second. It is, however, a very feeble oscillator and evidences its oscillations only by transient clicks heard in a telephone in the plate circuit as the tuning condenser passes through resonance.

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At *B* is shown a Nichrome oscillator of frequency 102,000 cycles per second in a sealed glass tube in vacuo, in which it is loose. In the picture, gravitation has pulled the oscillator to the lower end of the inclined glass tube. In operation the glass tube is placed axially in a horizontal position in the coils of the oscillator, and the rod is then centrally supported within the glass tube and within the coils. The mounting in a sealed glass tube has the advantage of permanence and avoids heating of the rod by the body temperature when it is handled. The absence of air or gas in the tube prevents the production of stationary sound waves in the tube, which might slightly affect the period of the rod, but this effect is very small.

TABLE I

LENGTH AND FREQUENCY OF STOIC METAL RODS AT 20 DEG.C. DIAMETER 0.79 CM.

Specimen No.	Frequency Cycles per sec.	Length Cm	Length in M × Frequency
10	10001	20.815	2081.7
11	11005	18.91	2081.0
12	11985	17.36	2080.6
13	12990	16.01	2079.7
14	14000	14.85	2079.4
15	14981	13.87	2077.8
16	16009	13.00	2081.2
17	17006	12.24	2081.5
18	18015	11.55	2080.7
19	19013	10.94	2080.0
20	20003	10.40	2080.3
21	21007	9.90	2079.7
22	22008	9.44	2077.5
23	22970	9.035	2075.2
24	24009	8.65	2076.8
25	24992	8.33	2081.8
26	25979	8.01	2080.9
27	26981	7.70	2077.5
28	27966	7.44	2080.5
29	29000	7.17	2079.2
30	29981	6.93	2077.7
		Average	2079.6±2

$v_{stoic} = 2 \times \text{Length} \times \text{Frequency} = 4160 \pm 2$ meters per sec. at 20 deg. C.

A great variety of rod sizes are in use in various researches now under way at this laboratory.

Lengths and Frequencies of Rods. Table I contains data obtained with a set of Stoic metal rods. These are the rods shown at *G* in the photograph. The frequency values given in the second column are accurate to 1/100 of 1 per cent. The lengths are not accurately known because the ends of the rods

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are not accurately true. The errors in the measurements of these lengths are of the order of 0.1 mm which amounts to more than 1/10 of 1 per cent for the shorter rods.

It is seen that the length of the rods times their frequencies gives a constant to the degree of precision to which the lengths are measured. This constant, when doubled, gives the velocity of sound in the metal, which is

$$v_{stoic} = 4160 \pm 2 \text{ m per sec. at 20 deg. C.}$$

All of these rods have roughly the same diameter, and the fact that there is no progressive change in the values of the last column indicates that there is no appreciable end-correction for the velocity determination.

The importance of this result for the present purpose is that a whole series of vibrators can be precomputed as to length with sufficient accuracy to make their final adjustment to required frequency very simple. A rod can be predetermined and cut in a lathe to 0.01 cm so that the final frequency adjustment to a standard value requires very little grinding. The end is ground off to raise the frequency. If too much is ground off at the end, it can be corrected by grinding away a little from the girth near the center of the rod.

Table II contains a similar set of observations of frequency for different lengths of Nichrome. Velocities for a large number of alloys are given below.

Experimental Data Illustrating the Manner in Which the Magnetostrictive Rod Functions in Controlling Frequency. Illustrative of the action of the rod as a frequency-controlling element the experimental curves of Fig. 2 are given. These are from observations taken with a rod (No. 17) of Nichrome in the coils and circuit of Fig. 1. In the lower part of Fig. 2 electric wavelengths λ, read on a precision wavemeter, are plotted against condenser divisions of the condenser C of Fig. 1. Readings were taken first with the rod restrained by holding it with the hand. A plot of λ versus C for the rod thus restrained is the curve $ABCDE$ of Fig. 2.

Next readings were taken with the rod free, starting with small values of the condenser C. These readings plotted give the curve $ABCD'E$, with a constant λ along the element D' of the path.

As a third operation, with the rod free the condenser was

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started at large values of condenser and gave the curve $EDB'A$, with constant λ along the element B' of the path.

As a fourth operation it is noted that when we have once got the rod vibrating by either of the above operations we can pass back and forth along the whole of B' and D' (i.e. from 18 divisions to 30 divisions of the condenser) and the wavelength remains essentially constant.

TABLE II

LENGTH AND FREQUENCY OF NICHROME RODS AT 23 DEG.C. DIAMETER 0.96 CM.

Specimen No.	Frequency Cycles per sec.	Length Cm	Length in M × Frequency
26	25999	9.58	2490.7
27	27000	9.21	2484.0
28	28000	8.87	2486.4
29	29005	8.56	2482.8
30	29992	8.27	2480.3
31	31005	8.02	2486.6
32	32011	7.77	2487.2
33	33006	7.55	2491.9
34	34011	7.32	2489.6
35	35002	7.13	2495.6
36	36002	6.88	(2476.9)*
37	37004	6.73	2490.3
38	38004	6.55	2489.2
39	39005	6.38	2488.5
40	40004	6.23	2492.2
41	41007	6.06	2485.0
42	42000	5.93	2490.6
43	43005	5.79	2490.0
44	44009	5.66	2490.9
45	45008	5.53	2488.9
46	46004	5.42	2493.4
47	47009	5.30	2491.4
48	48011	5.20	2496.5
49	49008	5.09	2494.5
50	49996	4.99	2494.8
51	51040	4.87	2485.6
52	52000	4.78	2485.6
53	53062	4.69	2488.6
54	54027	4.61	2490.6
55	55014	4.53	2492.1
56	56042	4.45	2493.9
57	57015	4.37	2491.5
		Average	2490.3

$v_{\text{Nichrome}} = 4981$ m per sec. at 23 deg. C.

*Inconsistent on account of grinding at girth for adjustment.

The wavemeter measurements are not sensitive enough to detect any change of frequency along this path $B'D'$. To measure this quantity resort was had to beats between the oscillating system with rod No. 17 under investigation and a second oscillator with magnetostriction-rod control. Since the total change of

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No. 17 was of the order of one cycle per second in seventeen thousand, the second oscillator, known as No. 117.7, was chosen to have a frequency about seven times that of No. 17, and beats were observed between the seventh harmonic of No. 17 and the fundamental of No. 117.7. This means that one cycle per second change in No. 17 gives seven cycles per second change in the beat note, which can be measured to about one cycle per second.

With this device the actual frequencies of the system stabilized by rod No. 17 against condenser settings are plotted as the top curve of Fig. 2. This curve shows that a change of the condenser from 19 to 30 divisions changes the frequency of the system by only 1.5 parts in seventeen thousand. This effect is discussed further in the next section.

In this experiment I have changed the condenser through a wide range of values, so as to make the change of frequency as much as possible, but when used as a constant frequency apparatus the condenser is brought always to the value which makes the plate current a maximum so that repeated settings of the apparatus may be made with a much higher degree of precision than the 1.5 cycles in seventeen thousand cycles per second.

Let us look next at the middle curves of this Fig. 2, which exhibits the plot of plate current against condenser settings, when the rod is free to vibrate. The arrows indicate the sense of a change given to the condenser of the oscillator circuit. When the condenser, on being increased, reaches 24.5 divisions the rod begins to vibrate and the plate direct current jumps from two to four milliamperes, and remains about the same for further increases of the condenser up to 30.1 divisions when the rod ceases to vibrate. Now a reduction of the condenser leaves the plate current two milliamperes till $C=25$ divisions is reached when the plate current jumps again to four milliamperes and then gradually decreases as the condenser is changed further down to eighteen divisions, when the rod is found to have again ceased to vibrate. In fact with this particular set of apparatus constants the system ceases also to vibrate electrically at C equal to or less than eighteen divisions. This accounts for the smooth fall of plate current to its constant two-milliampere value. With more clearance in the coil between the rod and the coil or with other electrical or mechanical constants we may get an abrupt change of plate current at the left-hand part of the curves as well as at the right-hand part. With less clearance, or by putting a separate

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condenser about each of the coils of Fig. 1, the apparatus can be so constructed that all oscillations cease whenever the rod is restrained from vibrating.

The Apparent Reactance of the Coil Containing the Tuned Magnetostriction Rod a Function of the Frequency. A crude view of the above result is that in the range of frequency-stabilization the system oscillates as a condenser-inductance circuit

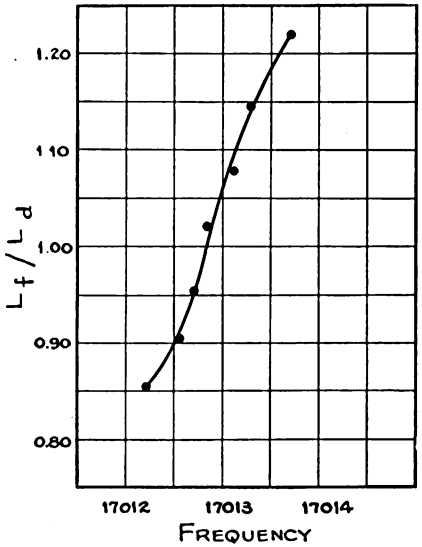

Fig. 3—L_f = Inductance Free; L_d = Inductance Damped; Plotted Against Frequency.

with an inductance that varies enormously with a small change of frequency. If we regard the wavelength-condenser curves of the lower part of Fig. 2 as having the equation

$$\lambda_d = 2\pi c\sqrt{L_d(C+C_o)} \text{ for the damped curve}$$

and

$$\lambda_f = 2\pi c\sqrt{L_f(C+C_o)} \text{ for the free curve}$$

we obtain

$$\left(\frac{\lambda_f}{\lambda_d}\right)^2 = \frac{L_f}{L_d}.$$

The values of this ratio are plotted against frequency in the curve of Fig. 3, from which it appears that the inductance of the

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coil and free rod, in the range of its stabilization frequencies, changes by about 50 per cent for a frequency of 1.5 in 17013; that is, for a frequency change of less than 1/100 of 1 per cent.

Though the correct understanding of the device requires mathematical discussion, such as is given below, we may here regard the constancy of frequency, with large variation of condenser, as due to the fact that an increase of C makes the frequency lower. This is accompanied by a shift of L along the curve of Fig. 3 to a smaller value so that the product of $L \times C$, and consequently the frequency, is almost the same as before the increase of C and, in fact, with this particular magnetostrictive material (Nichrome) the frequency changes only 1/100 of 1 per cent for a change of 50 per cent in the value of C. The potency of the rod in controlling the frequency is seen by noting that if we omit the rod or restrain it from vibrating this change of 50 per cent in C causes 25 per cent change in frequency instead of the 1/100 of 1 per cent with the rod present and free.

Frequency Essentially Independent of Vacuum-Tube Voltages and Characteristics. I have made numerous experiments that show that the very large change of plate voltage from 135 volts to 67 volts changes the frequency of oscillation of the magnetostriction controlled system by only about one in thirty thousand. Voltage change of the filament battery from a condition of practically zero emission to the point of destruction of the filament or a change of the inductance of driving coils over the whole range of oscillatory operativeness or a change of the tubes to tubes of widely different characteristics affects the frequency within about the same limits.

These various notes as to constancy apply to annealed Nichrome rods. Some of the other materials and particularly certain tubular vibrators are not quite so constant in these particulars, but are still very serviceable when less precision is required.

Temperature Coefficient of the Nichrome Vibrators. The temperature coefficient of the frequency of the Nichrome vibrator is

$$\frac{\Delta f}{f\Delta\theta} = -0.000107/\text{deg. C.}$$

where f = frequency, Δf = increase of frequency for a change of temperature of $\Delta\theta$ degrees Centigrade. This equation means that the frequency decreases about 1/93 of 1 per cent per degree Centigrade.

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The temperature coefficient of other materials is given in a later section.

Comparison of Constancy with That of the Piezo-Electric Crystal Vibrators. For the purpose of comparison of the magnetostriction oscillators with the piezo-electric crystal method of frequency control to which I have contributed in previous researches I have made extreme tests on some of my best piezo-electric crystal frequency standards. One is a crystal, No. 28, mounted *in vacuo* in an accurately designed and carefully machined holder. When operated under constant conditions its absolute frequency has been repeatedly measured and shows variations below 1/500 of 1 per cent. However, care must be taken to keep the circuits constant with this crystal oscillator as the data of Table III show.

TABLE III
PIEZO-ELECTRIC CRYSTAL OSCILLATOR

Conditions		Frequency
Tube	Plate Volts	
199	45	28072
Daven 20	67	28064
Daven 20	112	28061
Daven 20	135	28058
201A	90	28056
201A	112	28054
201A	135	28052

It is seen that extreme variations of tubes and plate voltage changes the frequency of this crystal oscillator by 7/100 of 1 per cent.

This crystal also changes frequency to about the same extent with change of the inductance and distributed capacity of the plate coil. Table III compared with previous sections of this account shows that for equally violent change of the circuits, tubes, and voltages with the piezo-crystal oscillator and the magnetostriction oscillator the magnetostriction oscillator is much more constant as to frequency than is the piezo-crystal oscillator.

It should be noted, however, that in the matter of temperature coefficient of frequency the particular Nichrome magnetostriction oscillator with 1/93 of 1 per cent per degree Centigrade is more subject to temperature changes than the piezo-electric oscillators with 1/200 to 1/1000 of 1 per cent per degree Centigrade.

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Below I give, however, descriptions of other magnetostrictive materials with temperature coefficients of about the same order as that of the piezo-electric crystal oscillators. The temperature effect in no case is permanent and the frequency comes back to its calibrated value when the temperature returns to normal for the calibration. A knowledge of the temperature permits the application of a correction to the rated calibration.

Magnetostriction Oscillator with Universal One-Stage Amplification. In order to increase the output power one or more stages of amplification may be employed. Fig. 4 shows a very simple and useful arrangement for a system to be operated at frequencies extending all over the audio- and radio-frequency

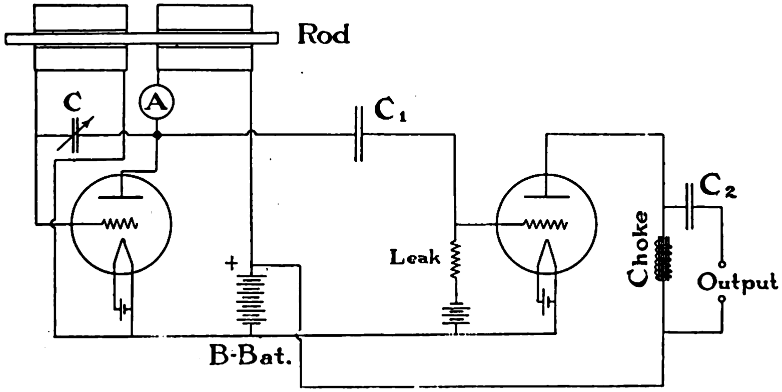

Fig. 4—Magnetostriction Oscillator with One-Stage Amplification. C_1 = 0.1 μf; C_2 = 2 μf; leak = 0.05 megohms; *A* = milliammeters.

ranges. The oscillator tube shown at the left has the connections of Fig. 1. The grid of the amplifier tube at the right is connected through a condenser C_1 to the plate of the oscillator tube, and is provided with a leak resistance. The plate circuit of the amplifier tube contains a "speaker filter" unit consisting of an audio choke which is in series with the B-Battery and which communicates with the output terminals through a condenser C_2 of two microfarads.

The constants given in the legend to Fig. 4 are so chosen that the amplifier operates with considerable efficiency at both radio and audio frequency. In the latter case a loudspeaker connected to the output terminals gives volume enough to be heard throughout a large lecture room. This is the case also with radio oscilla-

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tions upon which are superposed other radio oscillations to produce audio-frequency beats.

A commercial apparatus of this character made for me by the General Radio Company of Cambridge is shown in Fig. 15, and at the left in Fig 16. The coil units, containing rods, on top of the boxes are plugged into jacks and can be removed and interchanged. The upper cover of the box opens.

Twin Magnetostriction Oscillator. For calibration purposes we need two oscillators like that of Fig. 4, one of which is maintained constant by a standard magnetostriction rod, and the other of which is at a frequency to be compared with the rod. The two oscillators are then caused to interact and their beat frequency is determined or reduced to zero. After much experimenting I find that the best way to get the interaction is to connect their output terminals together through a small condenser and then to plug the telephone or other indicator in at one or the other of the output terminals. If the rod is left out of one of the coil units that half of the twin oscillator may be used as an ordinary variable electric oscillator.

I have found so much use for this arrangement that I have had both of these oscillators with common B-batteries and common A-battery terminals mounted in a single apparatus shown as the middle unit of Fig. 16.

The "speaker filters" of the two output circuits are connected with a switch so that if desired, either or both may be cut out and the B-battery and plate of the amplifiers be connected directly to the output terminals. This is useful in running a synchronous-motor clock, which requires a polarizing current, as the output apparatus. Such a clock made also by the General Radio Company from their own designs and kindly put at my disposal is shown in Fig. 15, and was used in timing some of the low-frequency rods (at about 1000).

Absolute Frequency Calibration by Magnetostriction Oscillators and the General Radio Synchronous-Motor Clock. Use is made of the adjustable magnetostriction rod of about 1000 cycles per second, which is shown in Fig. 15 mounted on top of the box containing the electric oscillator circuits and the one-stage universal amplifier. Used with the "speaker filter" cut out this drives the motor clock synchronously with the oscillations of the rod. The amplifier tube is a UX171, operating on 130 volts in the plate.

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The synchronous clock has been provided with circuit-making contacts that close for a very short time each second of the clock.

For a standard of time with which to compare this motor clock Professor Stetson, of the Harvard Students' Astronomical Laboratory, has kindly permitted me to lead wires from his standard Riefler clock, which beats siderial seconds with an error not greater than two-tenths of a second per week. All relay contacting members are obviated in this circuit by the device of placing an inductance coil near the field coil of his relay, so that the local

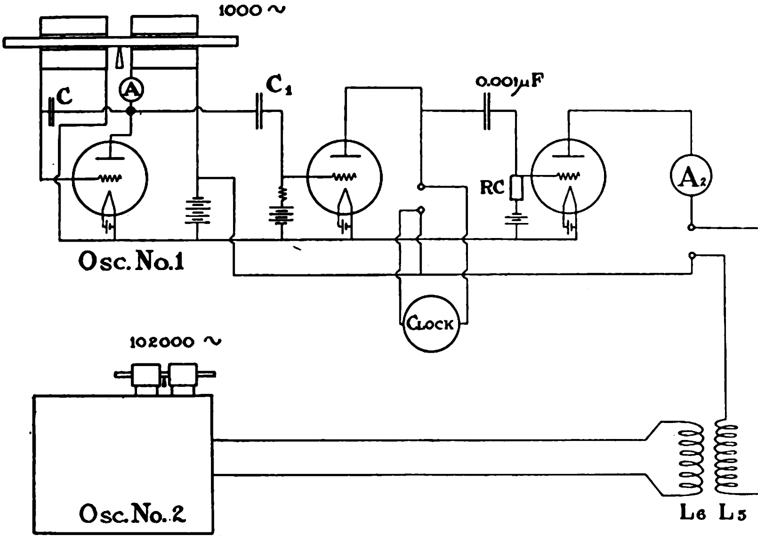

Fig. 5—Circuits for Absolute Calibration of Frequency.

current from the time wheel of the clock through his relay field acts inductively on my circuit and supplies second impulses without the inconstancy of any contact other than that of the time wheel of the clock. These second impulses arriving at this laboratory are amplified by a one-stage amplifier, to the output of which is connected a loudspeaker through the contacting points of the motor driven clock. When the standard Riefler clock and the motor clock make contacts at the same time the ticks are heard in the loudspeaker. By reading the motor clock for several successive coincidences accurate timing is obtained of the motor clock and of the magnetostriction driving rod.

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For timing any other suitable magnetostriction rod, (say one of 102,000 vibrations per second) the latter is used to drive and control a second oscillator mounted on a second box like that of Fig. 15 or else the twin system is used with both the rods. The 1,000-cycle rod is now adjusted by moving the weights out or in along the rod to points where zero beat is obtained between the 102nd harmonic of the 1,000-cycle rod and the fundamental of the 102,000-cycle rod. When this adjustment is nearly perfect the remaining small adjustment is made by changing the tuning condenser of the 1,000-cycle system. In this way I have succeeded in reducing the beat frequency to one cycle in five seconds, which means an error of this adjustment of the harmonic of the low frequency to the fundamental of the high frequency of two parts in a million. Such accuracy is ordinarily not required, so that more tolerance is usually given to this adjustment.

It is interesting to note that the beats may be observed on a d. c. milliammeter as well as in a telephone. As a means of obtaining beats on the milliammeter the circuits of Fig. 5 are recommended.

In Fig. 5 Oscillator No. 2 has the circuits of Fig. 4 with a coil L_6 connected to its output terminals. Oscillator No. 1 has the same circuits for the two left-hand tubes with, however, the motor clock in place at the choke of the second tube. Paralleling the clock a radio-amplifying unit is employed for further amplifying the higher harmonics of Oscillator No. 1 and for suppressing its 1,000-cycle note and low pitch harmonics. The connection to the third tube is through a condenser of 0.001 μf, which has high impedance to these low notes. The customary grid leak of this third tube is replaced by a radio choke *R C* of about 0.1 henry, which has a small impedance for the low notes and a high impedance for the higher harmonics. The output circuit leads to L_5 inductively connected with L_6 so that the oscillations of the two systems are superposed in their output circuits, and their beats when nearly of zero frequency are observed on the d. c. milliammeter A_2.

In this way timing of the motor clock gives a direct standardization of the 102,000-cycle rod as well as of the 1,000-cycle. rod.

In a similar way any of the rods may be calibrated directly in one step. I have used harmonics of the 1,000-cycle rod up to its 234th.

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To calibrate at the same time directly any higher frequency a double step is required.

Examples of Absolute Calibration of Some of the Rods. In calibrating Nichrome rod No. 30, the rod No. 1 (of about 1,000), was adjusted to give zero beats of its 30th harmonic with rod No. 30. Coincidences of the rod-clock seconds with the Riefler clock siderial seconds occurred as follows:

Coincidences Read on Rod Clock			Time Between Coincidences in Rod-Clock Seconds
hr.	min.	sec.	
3	14	31	396
3	21	17	395
3	27	42	399
3	34	21	

For computation we use

$$1 \text{ solar true second} = 1.00273 \text{ siderial seconds.}$$

$$1 \text{ rod second} = 1 + \frac{1}{397} = 1.00252 \text{ siderial seconds.}$$

$$\therefore \quad 1 \text{ solar second} = \frac{1.00273}{1.00252} = 1.00021 \text{ rod seconds.}$$

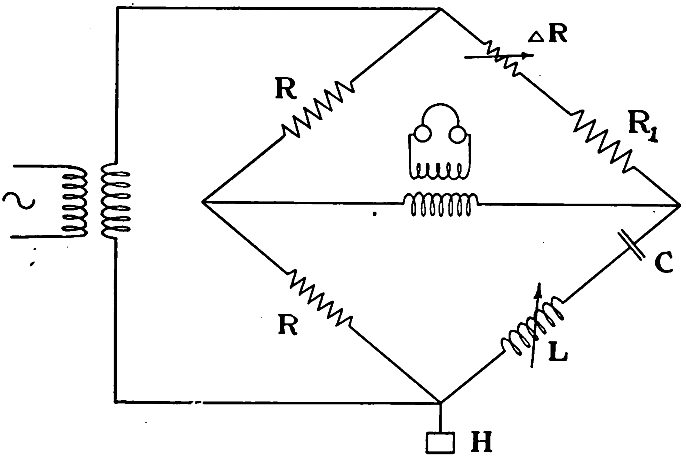

Fig. 6—Circuits of Audio-Frequency Meter.

The motor clock operated by the rod has its hands so geared to its motor that its time is correct when the motor is driven by a current of 1,000 cycles per second, so we have for the frequencies of the rods:

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$$f_{\text{no. }1}=1000.21,\ f_{\text{no. }30}=30{,}006.3 \text{ at } 20.4 \text{ deg. C.}$$

Two other independent determinations of this rod No. 30 give 30,007.2 and 30,006.9 at 20.2 deg. C. The average of the three values is 30,006.8, with an average error of ½ cycle. Each of these calibrations had a timing period of about 20 minutes; by prolonging the time any required greater accuracy can be obtained.

The following values, Table IV, taken at random from my notebook have been obtained in the absolute calibration of a number of standards, and illustrate what can be done with even very careless work provided the standard second is good and is applied sharply to the apparatus.

TABLE IV
SAMPLE ABSOLUTE CALIBRATION OF STANDARD

Standard No.	Rod No. 30 at 20.3 deg. C.	Rod No. 102 at 20.6 deg. C.	Rod No. 117.7	Crystal No. 5884
Cycles per Second	30006.3 30006.9 30007.2	102078 102085 102084 102084 102083 102084	117718 at 21.9 deg. C. 117679 at 24.5 deg C 117705 at 23.0 deg C.	5884707 5884722
Average	30006.8	102083	117701 at 23 deg. C.	5884719

Referring to the values for rod No. 117.7 the three values were taken at different temperatures and if reduced to a common temperature of, say 23 deg., these values become 117704, 117689, 117705, averaging 117701 with an average error of 2 cycles per second. With this correction for temperature the values for No. 117.7 have about the same consistency as the other values in the table, and the variations in the table amounting to about 1/500 of 1 per cent are due to inaccuracy arising from the paucity of the coincidence observations.

The measurements of the frequency of Crystal No. 5884 of about 5884 kilocycles were obtained by using the rod of 1000 cycles adjusted to 997.28 cycles so that its 118th harmonic beats zero with rod No. 117.7 and at the same time beats were measured between the 50th harmonic of No. 117.7 and the fundamental of Crystal No. 5884. The latter number of beats per second was found to be 772, so we have $f_{5884}=50(118\times 997.28)+772=$ 5,884,722. This double operation steps up the 1000-cycle frequency to its 5900th harmonic.

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It is interesting to note that this crystal No. 5884 can also be used at all of its harmonics up to its 30th harmonic to measure frequencies up to 150 million cycles.

Mr. M. T. Dow, Dr. Frederick Drake, and I are using such a calibration in the precision measurement of the velocity of electric wave.

By the methods here described the several rods of Fig. 14 have been calibrated and intercompared. In the intercomparison the differences between fundamentals or harmonics of various oscillators may be conveniently measured by a bridge-type of audio-frequency meter such as is described in the next section.

Audio-Frequency Meter. I here call attention to a bridge-type of audio-frequency meter involving well known circuits[1] that I have found useful for measuring audio frequencies. A photographic view of this apparatus is shown as the right-hand unit of Fig. 16. The circuits are shown in Fig. 6 and comprise a bridge of three resistance arms and one arm of a variable L in series with fixed C. The two resistances R R are equal. The resistance R_1 plus ΔR is equal to the resistance of the branch LC which is chiefly the resistance of L. Since L is of copper wire it has a temperature coefficient. This is compensated by the small adjustable resistance ΔR, which when necessary is independently adjusted when the bridge is to be balanced. The chief variable element is L. At any given frequency, f, within the range of the apparatus the balance is obtained when

$$2\pi f = \frac{1}{\sqrt{LC}} \cdot$$

Frequencies are read on a direct-reading scale calibrated in frequency attached to the dial of L. One important point is that for repeatable readings with an accuracy of 1 part in 1000 the variable effects of body-capacity must be eliminated. I have found it advisable to provide the inductance L with a metal knob so that when one takes hold of this knob to adjust L he connects his body to the point H in Fig. 6.

The inductance L at full scale is about 200 millihenries. The capacity C is a three-step mica condenser, having the three values 0.04, 0.2, and 1.0 μf, which gives the instrument a three-

[1] Although this instrument has been in use here for many years it was apparently previously described and used as a frequency meter by Heydweiller and Hagemeister: *Verh. d. D. Phys. Ges.* 18, 52, 1916. For other references see Banneitz: *Taschenbuch der drahtlosen Tel. und Tel.*

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fold scale extending from 365 to 5000 cycles per second. The switch in the center of the photographic view of the instrument is for switching to different values of the condenser, and automatically switches the pointer to the proper direct-reading scale. The scale was drawn by hand on a bakelite disc from experimentally determined frequency readings and was then engraved on the disc.

In the instrument are included an input transformer, for connecting directly into the input circuit of a vacuum-tube oscillator, and an output transformer connecting to head telephones or a loudspeaker.

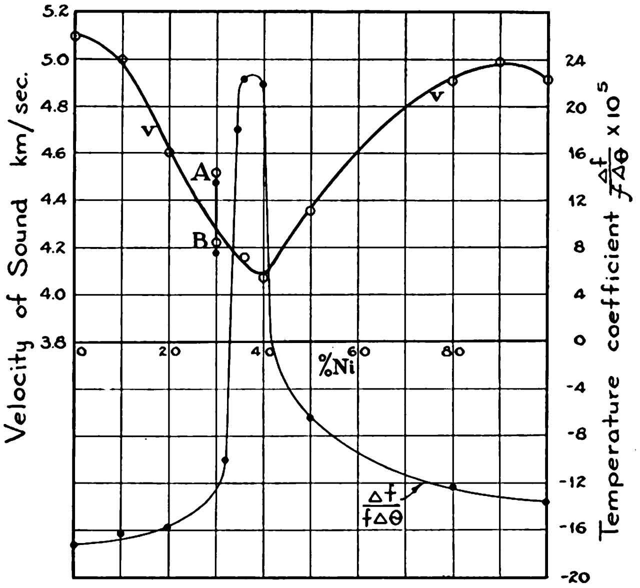

Fig. 7—Velocity and Temperature Coefficient of Frequency of Nickel-Iron Alloys of Various Percentages of Nickel.

Absolute Calibration of Audio-Frequency Meter of Fig. 6 by Nichrome Rod No. 30 with $f = 30{,}000$. An electric oscillator with frequencies variable to cover the range of the audiometer is made to beat aginst various harmonics of the fixed 30,000-cycle frequency of the rod No. 30. Practically, in the present calibration, this was done by using the Twin Oscillator of Fig. 16 (shown at the center) with the magnetostriction rod in its coil assembly plugged into one set of jacks on the instrument and with various coil-

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pairs without magnetostriction core plugged into the other set of jacks. One pair of the output terminals of the Twin Oscillator were now connected to the input posts of the audio-frequency meter to be calibrated. A telephone receiver was at the output posts of the audio-frequency meter.

The procedure is as follows: Adjust the variable electric oscillator to beat zero with some harmonic of the magnetostriction rod, then adjust the audio-frequency meter for silence in the telephone and read the scale of the audio-frequency meter. Proceed *ad lib.* with other harmonics.

Table V gives the results. The value of the harmonic used is given in the column marked n. For example $n = 3/50$ means a beat zero between the 50th harmonic of the 30,000 cycles rod and the 3rd harmonic of the electric oscillator, to which the audiometer was set, giving $f = 1800$ cycles per second. This is tabulated in the second column.

The observations in this Table V were made hurriedly without any repetition of readings. Only the more striking harmonics (simple ratios) were used, and a large number of other values were passed over without being read, as there was no need of obtaining any more dense distribution of the calibration points. The accuracy of reading the scale of the instrument was only about 1 part in 1000 in the more open parts of the scale. Checks on the identity of the harmonics can be obtained by using other magnetostriction rods.

Absolute Calibration of a Wavemeter by Nichrome Rod No. 30. To show a further use of the magnetostriction standards this same rod of 30,000 cycles per second ($\lambda = 10,000$ meters) was used to obtain the values of Table VI in calibration of a wavemeter ranging from 233 to 2224 meters. The procedure was the same as in the preceding section, except that the electric oscillator coils were of small inductance so as to give high frequency. The telephone connected directly at the output of the Twin Oscillator served to evidence audible beats, which were adjusted to a zero-beat-frequency.

The wavemeter was placed near the electric oscillator coil and had its coil shortcircuited, while setting the oscillator to beat zero. Then the shorting switch of the wavemeter was opened and the condenser of the wavemeter was adjusted. As resonance was approached beats were again heard in the telephone. Changing the condenser of the wavemeter in one direction, we hear

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these beats rise in frequency, then drop to zero, then appear again at a high frequency and gradually change to zero. The drop of the beats to zero at the setting between the two high-frequency beat values is the adjustment of the wavementer to resonance.

In Table VI n indicates the harmonics employed. This is here tabulated as $1/n$, since we are concerned with wavelengths. The wavelength λ of the wavemeter is obtained by taking $1/n$ times the electric wavelength (10,000) of the 30,000-cycle rod.

TABLE V
CALIBRATION OF AUDIO-FREQUENCY METER VS. NICHROME ROD NO. 30.

Scale of Instrument	f by No. 30	n	Scale of Instrument	f by No. 30	n
872	870	2/69	1760	1767	1/17
897	896	2/67	1800	1800	3/50
909	909	1/33	1818	1818	2/33
922	923	2/65	1871	1875	1/16
937	938	1/32	1909	1915	3/47
952	952	2/63	1935	1935	2/31
956	957	3/94	2000	2000	1/15
968	968	1/31	2068	2069	2/29
978	978	3/92	2140	2143	1/14
982	984	2/61	2223	2222	2/27
1090	1091	2/55	2305	2308	1/13
1130	1132	2/52	2395	2400	2/25
1181	1176	2/51	2500	2500	1/12
1228	1224	2/49	2605	2609	2/23
1252	1250	1/24	2722	2727	1/11
1278	1277	2/47	2999	3000	1/10
1330	1333	2/45	3165	3158	2/19
1390	1395	2/43	3330	3333	1/9
1462	1463	2/41	3535	3529	2/17
1498	1500	1/20	3740	3750	1/8
1543	1538	2/39	4000	4000	2/15
1581	1579	1/19	4285	4285	1/7
1603	1607	1/23	4478	4500	3/20
1620	1622	2/37	4595	4615	2/13
1660	1666	1/18	4710	4737	3/19
1692	1696	3/53	5010	5000	1/6
1710	1714	2/35			

In this section and the preceding section it is seen that the 30,000-cycle rod is used in frequency calibrations extending from 870 cycles to 1,200,000 cycles. The possible range is vastly greater than this.

In practice I no longer use directly the wavemeter or the audio-frequency meter for accurately measuring wavelengths or frequencies, but instead beat everything against the rods or piezo-crystals, and use the audio-frequency meter for measuring

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differences of frequencies, and the wavemeter for identifying harmonics.

Study of Frequency, Sound Velocity, and Temperature Coefficients of Frèquency of Various Alloys. I have investigated a large number of alloys for their magnetostrictive properties, and,

TABLE VI

CALIBRATION OF WAVEMETER LC_1 VS. NICHROME ROD NO. 30. READINGS ARE IN METERS WAVELENGTHS

Scale of Instrument	λ by No. 30	1/*n*	Scale of Instrument	λ by No. 30	1/*n*
2224	2222	2/9	866	869	2/23
2175	2174	5/23	832	825	1/12
2144	2143	3/4	799	800	2/25
2106	2105	4/19	768	769	1/13
2085	2084	5/24	712	714	1/14
2070	2068	6/29	662	666	1/15
2000	2000	1/5	621	625	1/16
1940	1944	7/36	574	578	1/17
1924	1923	5/26	553	555	1/18
1892	1892	7/37	523	526	1/19
1870	1875	3/16	514	513	2/39
1848	1852	5/27	502	500	1/20
1815	1818	2/11	475	476	1/21
1790	1786	5/28	454	455	1/22
1712	1714	6/35	435	435	1/23
1668	1667	1/6	416	417	1/24
1573	1579	3/19	401	400	1/25
1535	1538	2/13	383	385	1/26
1510	1515	5/33	368	370	1/27
1500	1500	3/20	356	357	1/28
1480	1481	4/27	344	345	1/29
1462	1463	6/41	333	333	1/30
1426	1429	1/7	322	323	1/31
1388	1389	5/36	312	313	1/32
1378	1379	4/29	303	303	1/33
1362	1364	3/22	295	294	1/34
1350	1352	5/37	285	286	1/35
1332	1333	2/15	276	278	1/36
1303	1304	3/23	270	270	1/37
1251	1250	1/8	263	263	1/38
1201	1200	3/25	257	256	1/39
1178	1176	2/17	251	250	1/40
1111	1111	1/9	245	243	1/41
1050	1052	2/19	238	238	1/42
999	1000	1/10	233	233	1/43
952	952	2/21			

by measuring the frequency of oscillation of a known length of specimen, have determined the velocity of a longitudinal sound wave in these various materials. These are recorded in column 2, Table VII. By making these measurements at various temperatures I have obtained the temperature coefficient of fre-

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quency (column 3). By the formula

$$v=\sqrt{\frac{\text{Young's modulus}}{\text{Density}}}$$

I have computed Young's modulus (column 5) from measured values of velocity and density (column 4). This method gives high precision, and is interesting in that the value of the modulus is obtained for very small applied forces. Additional data of Table VII are described below.

TABLE VII

Experimental Results for Alloys

Material	v m/sec	θ $\times10^6$	ρ g/cc	$E\times10^{11}$ Dyne / sq. cm.	a $\times10^6$	h $\times10^6$	b $\times10^6$
Iron	5074	−171	7.688	19.79	12	−159	−354
Nickel	4937	−132	8.803	21.46	12	−120	−276
Stoic	4161	+224	8.02	13.89	1.5	+226	+446
Nichrome	4981	−107	8.269	20.52	12	− 97	−226
Monel	4549	−151	8.854	18.32	14	−137	−316
Stainless Steel	5430	−136	7.720	22.76	10	−135	−282
Stainless Iron	5133	−130	7.743	20.40	10	−130	−270
Nickel-Iron % Ni							
0	5074	−171	7.688	19.79	13.3	−158	−355
10	4919	−164	7.812	18.90	9	−155	−337
20	4582	−159	7.818	16.41	8	−151	−326
30	4527	+135			9	+144	+261
30	4241*	+ 77*	7.950*	14.29*	4*	+ 81*	+150*
32	4540	−101			3	− 98	−205
34		+182			2	+184	+362
36	4161	+224	8.020	13.89	1.5	+226	+446
40	4075	+218	8.042	13.36	5	+223	+431
50	4352	− 64	7.901	14.96	10	− 54	−138
60	non oscillatory						
70	non oscillatory						
80	4908	−124	8.521	20.52	12	−112	−260
90	4990		8.666				
100	4937	−132	8.803	21.46	12	−120	−276
Chrom-Iron % Cr.							
10	5290	−153	7.619	21.32	10	−143	−316
20	5448	− 90	8.028	23.82	10	− 80	−190
30	5392	−102	7.659	22.27	10	− 92	−214
40	4329	− 91	7.514	21.33	10	− 81	−192
Si Cr Fe%							
5 5 90	5166	−118	7.559	20.17			
5 10 85	5285	−111	7.581	21.17			
5 15 80	5192	−112	7.482	20.17			
5 20 75	4806	−144	7.488	17.30			
5 25 70	5473	−140	7.405	22.18			
5 30 65	5387	−133	7.414	21.52			
5.16 4.53 90		− 82					
Carbon-Steel % C							
0.8	5233	−110	7.849	21.49	11	− 90	−231
1.0	5209	−137	7.827	21.24	11	−126	−285
1.5	5217	−123	7.866	21.41	11	−112	−257

v = velocity of sound, θ = temperature coefficient of frequency, ρ = density, E = Young's Modulus, a = coefficient of expansion, h = temperature coefficient of velocity, b = temperature coefficient of elasticity.

* After dipping in liquid air.

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Other alloys are under investigation and will be reported in a later publication. Many of the specimens of the table contain a small quantity of manganese (about 0.3 per cent) added to make the specimen malleable.

Curves Showing Results for Velocity and Temperature Coefficients of Frequency with Alloys of Various Compositions. Fig. 7 shows velocity of sound in km per second and also temperature coefficient of frequency of vibration with rods of nickel-steel plotted against percentages of nickel of the rods. It is seen that the velocity has a minimum and the temperature coefficient has a maxi-

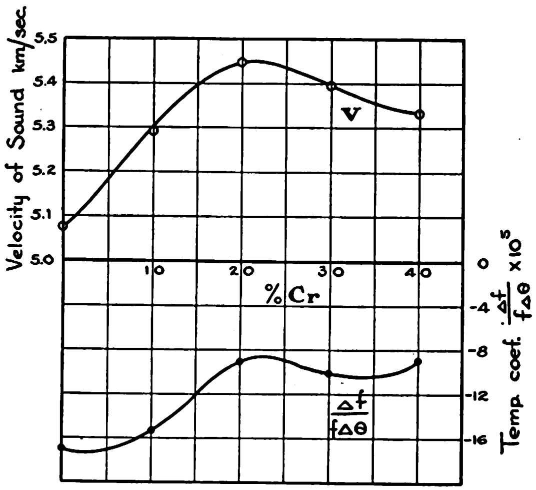

Fig. 8—Velocity and Temperature Coefficient of Frequency of Chromium-Iron Alloys with Various Percentages of Chromium (Abscissas).

mum at 36 per cent to 40 per cent nickel. This is at about the composition of Invar (36 per cent nickel). Fig. 8 contains corresponding curves for chromium-steel of various percentages of chromium; Fig. 9, similar results for various percentages of chromium in chromium-steels which contain also 5 per cent silicon. In Fig. 8 the velocity is a maximum and the temperature coefficient of frequency is a maximum at 20 per cent to 25 per cent chromium. These results, Fig. 7 and Fig. 8, suggest as a possible law *that in a binary metallic alloy the temperature coefficient of frequency is a maximum or minimum at the composition at which the velocity is a maximum or minimum, or the reverse.* Having found a means of extending the methods of the present research to non-magnetic

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alloys also I have assigned to a research student the problem of testing this apparent law for a wide variety of alloys.

The curves of Fig. 9 show the effects of adding a third component, 5 per cent silicon, to the chromium-iron series. This added silicon introduces a large minimum of velocity at 20 per cent chromium, 5 per cent silicon, and 75 per cent iron. Indications of corresponding maxima or minima of the temperature coefficient of frequencies at compositions giving a maxima or minima of velocity are also apparent in these curves, but the

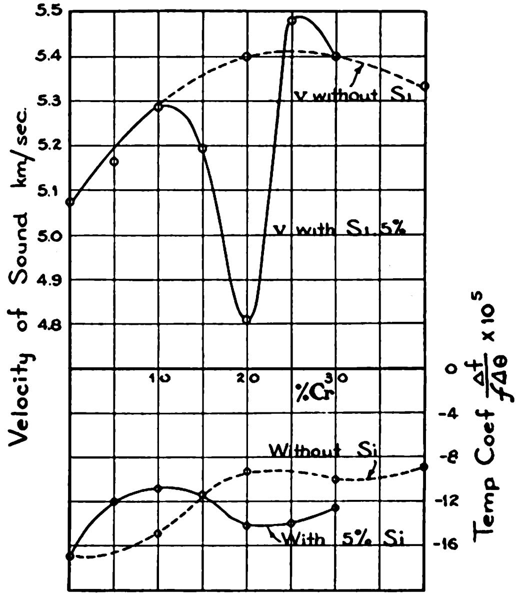

Fig. 9—Effect of Addition of 5 per cent Silicon to Chromium-Iron Alloys of Various Percentages of Chromium (Abscissas) and Rest Iron.

compositions employed were not selected closely enough together to make the result certain.

An interesting result is the largeness of the change of velocity with change of composition.

The points marked *A* and *B* of Fig. 7 were obtained with a nickel-steel of 30 per cent nickel. The values at *A* were obtained with the annealed specimen, and the interesting fact was noted that when the specimen was heated to about 50 deg. C it became non-magnetic and ceased to oscillate. When it was cooled down to 20 deg. C it again became magnetic and oscillatory, but re-

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versibly lost or gained its magnetism as the 50 deg. C point was passed. This fact for about this composition has been previously observed by Hopkinson.[2] Acting on the facts there given regarding effects of cooling, I had my specimen dipped into liquid air. After coming back to room temperature it gave the points marked *B* in Fig. 7. It had become a new substance and no longer lost its magnetism with change of temperature over the range up to 100 deg. C.

Temperature Coefficients of Elasticity and of Velocity. Simple relations exist among various temperature coefficients, so that other coefficients may be obtained from the experimental data of Table VII combined with known quantities taken from the work of previous investigators.

Given $\Delta\Theta$ = increment of temperature,

$$a=\frac{\Delta l}{l\Delta\theta}=\text{coefficient of linear expansion,}$$

$$b=\frac{\Delta E}{E\Delta\theta}=\text{coefficient of Young's modulus } E,$$

$$g=\frac{\Delta f}{f\Delta\theta}=\text{coefficient of frequency of longitudinal vibration,}$$

$$h=\frac{\Delta v}{v\Delta\theta}=\text{coefficient of velocity of longitudinal wave,}$$

and

$$2lf=v=\sqrt{\frac{E}{\rho}}=\sqrt{\frac{EV}{W}}\cdot \tag{1}$$

In this equation (1)

l = length of a specimen, which has
f = frequency,
ρ = density,
E = Young's modulus,
W = weight,
V = volume.

If the specimen satisfying equation (1) is raised 1 deg. C in temperature W does not change and we obtain

$$2lf(1+a)(1+g)=v(1+h)=\sqrt{\frac{EV}{W}(1+b)(1+3a)}$$

[2] See Ewing: "Magnetic Induction in Iron and Other Metals," Van Nostrand, 1900.

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Dividing this by (1) and performing the indicated operations for small values of the coefficients we obtain

$$1+a+g=1+h=1+\frac{b}{2}+\frac{3a}{2}$$

$$\therefore \qquad h=a+g, \text{ and } b=2g-a. \qquad (2)$$

Equation (2) enables us to compute the temperature coefficient of elasticity b and the temperature coefficient of velocity h from

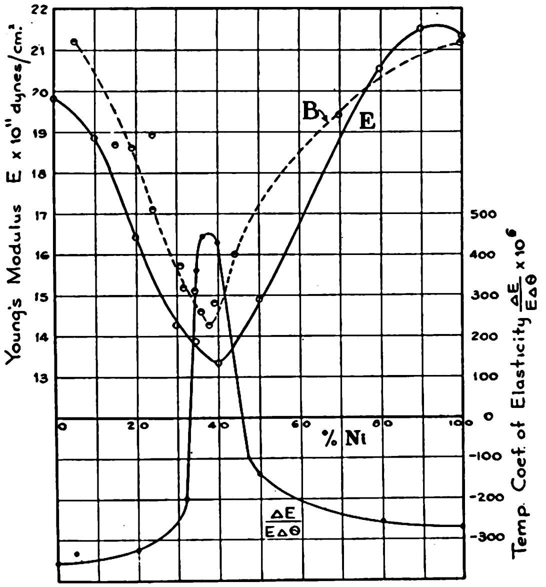

Fig. 10—Young's Modulus and Temperature Coefficient of Elasticity of Nickel-Iron Alloys with Various Percentages of Nickel (Abscissas).

the measured values of the temperature coefficient of frequency g and values of temperature coefficient of expansion a. Since the latter are small in comparison with g they are given sufficiently accurately in ordinary tables of constants.

Values of a taken from Bureau of Standards Circular No. 58 and from International Critical Tables are entered as column 6 of Table VII, and computed values of h and b are entered in columns 7 and 8.

Fig. 10 is a plot of E from my data on sound and also a plot of the corresponding values (Curve B) from the Bureau of

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Standards Circular No. 58. The large differences between these curves may be due to the fact that the values given in the Bureau Circular are obtained at large stresses while my values are for the very small stresses of sound waves. On the other hand, small differences in composition, as the 0.3 per cent manganese in my specimens, or different heat treatment, may account for the differences. It should be further noted that my specimens were magnetized.

Methods of Obtaining Magnetostriction Vibrators of Zero or Very Small Temperature Coefficients of Frequency. In Table VII the temperature coefficients of frequency g of magnetostriction vibrators of various alloys are given in column 3. In particular it is seen that the nickel-iron alloys vary from

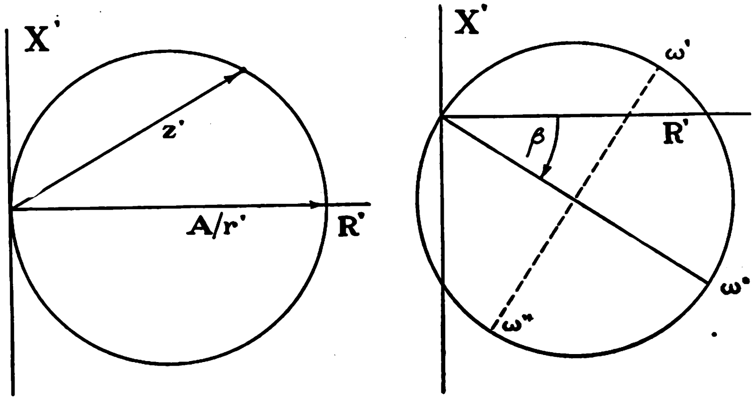

Fig. 11—Impedance Circle Diagrams.

$$g = -171 \times 10^{-6} \text{ to } g = +224 \times 10^{-6}$$

and that this coefficient is zero for alloys of 33 per cent Ni and 47 per cent Ni. Unfortunately these alloys are difficultt o produce with exact properties because the curve of temperature coefficient is very steep at these points. Furthermore, these materials are only very weak magnetostriction oscillators. With sufficient care in their production and heat treatment they may prove highly serviceable as oscillators of zero temperature coefficients. A further study of alloys of this class is under way.

In combinations of three metals I have found that a vanishing temperature coefficient is usually associated with non-magnetism or non-magnetostriction.

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Consequently I have investigated the alternative method of composite vibrators made of different metals so arranged that the positive coefficient of one specimen counteracts the negative coefficient of another specimen, which have the merit of simplicity of construction and can be made of materials giving a strong resultant vibration.

Two kinds of composite vibrators have been employed with success:

(*a*) A central rod of negative coefficient with a rod of positive coefficient soldered end to end at each end of the central rod. This I shall call *longitudinally composite.*

(*b*) A tube of coefficient of one sign containing a tight-fitting core rod of coefficient of the opposite sign. This I call *concentrically composite.*

A number of vibrators of both these types have been made. For standards the concentrically composite type have more power in controlling frequency and are preferred. Two such vibrators consisting of nickel tubes with stoic metal cores driven into them at a tight press-fit were found to have coefficients

$$\frac{\Delta f}{f\Delta\theta} = -\frac{1}{50000} \text{ and } -\frac{1}{65000}$$

respectively. Smaller values of this coefficient can be obtained by better choice of relative diameters of the components.

It should be noted that the cause of the change of any of these vibrators with change of temperature is the temperature coefficient of elasticity of the materials and not the change of length or change of density, both of which are relatively small.

Elementary Theory of Dynamic Magnetostriction. As a simplification of the more exact treatment of the subject, such as is given in Appendix I, there is here presented a discussion based on the assumption that the rod acts mechanically as a pair of masses separated by an elastic member and damped by friction proportional to velocity, so as to satisfy the equation

$$f' = m\ddot{x} + r\dot{x} + \epsilon Sx/l. \qquad (1)$$

In this equation the dots indicate time derivatives, and r, m, and ϵ are equivalent values of frictional resistance, mass, and Young's modulus respectively,

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S = Area of cross section

l = half length of the rod

x = increase of length of the half rod-length due to the force f'.

The force f' is a periodic magnetostrictive force due to a periodic magnetic induction B of comparatively small amplitude superposed on a fixed induction B_o, which is used to polarize the specimen.

Law I of Magnetostriction.—As a first law of magnetostriction we shall assume that the magnetostrictive force f' caused by the periodic induction B is proportional to B and to the area of cross section S; that is,

$$f' = aBS, \tag{2}$$

where a (positive for some substances and negative for others) is a coefficient depending on B_o but otherwise assumed constant.

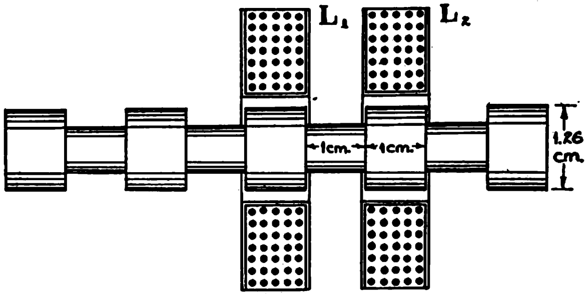

Fig. 12—Beaded Rods.

We shall regard B as made up of two parts B_i and B'

$$B = B_i + B', \tag{3}$$

where B_i = induction due to the periodic current i,

B' = induction due to the reverse magnetostriction effect (Law II).

Law II of Magnetostriction.—When the rod is stretched an amount $2x$ with an original rod-length $2l$ there is produced an induction B' proportional to x/l; that is,

$$B' = a'x/l. \tag{4}$$

Solution for Time-Derivative of x.—The substitution of (2), (3), and (4) into (1) gives

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$$aSB_i = m\ddot{x} + rx + \{\epsilon - aa'\}x/l, \quad (5)$$

of which the solution for the time-derivative of x on the assumption that B_i involves the time only in a factor of the form $e^{j\omega t}$ is

$$\dot{x} = \frac{aSB_i}{r + j\{m\omega - \epsilon' S/l\omega\}} \quad (6)$$

where as an abbreviation

$$\epsilon' = \epsilon - aa'. \quad (7)$$

Electrical Equations.—We shall next write the electrical equations. The alternating current i is related to the impressed emf e by the equation

$$e = Ri + \frac{d}{dt}\{\text{Flux linkage of } (B_i + B')\}. \quad (8)$$

The first two terms of the right-hand side of this equation give the electromotive force if the rod were damped so that it did not move. The remaining term gives the electromotive force due to the motion, which we shall call *motional* emf and shall designate by e', so

$$e' = \frac{d\phi}{dt}, \text{ where} \quad (9)$$

$$e' = \text{damped emf},$$

$$\phi = \text{Flux linkage of } B'.$$

If now we write

$$\phi = a''B' \text{ and} \quad (10)$$

$$B_i = L'i, \quad (11)$$

where a'' and L' are constants depending on the size, shape, and number of turns of the coil and on the permeability of the specimen, and combine (4), (6), (9), (10), and (11), we obtain

$$e' = z'i, \text{ where} \quad (12)$$

$$z' = \frac{aa'a''L'S/l}{r + j\{m\omega - \epsilon' S/l\omega\}}. \quad (13)$$

In equation (13) z' is the complex motional impedance of the system. ω is the angular velocity of the current. a, a', a'', and L' are coefficients defined by (2), (4), (10), and (11), respectively.[3]

[3] This theory and the discussion of the next section is essentially of the same character as that given by Professor Kennelly and the author for the impedance of telephone receivers. *Proc.* Amer. Acad. 48, No. 6, 1912, and *Electrical World*, September, 1912.

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Impedance Circle Diagrams. If we let

R' = motional resistance

X' = motional reactance

of the rod-coil unit, and as further abbreviations, write

$$A = aa'a''L'S/l \text{ and} \tag{14}$$

$$Z_m^2 = r^2 + \{m\omega - \epsilon' S/l\omega\}^2, \tag{15}$$

we find that (13) after rationalization gives

$$R' = \frac{Ar}{Z_m^2} \tag{16}$$

$$X' = \frac{-A\{m\omega - \epsilon' S/l\omega\}}{Z_m^2} \tag{17}$$

whence

$$R'^2 + X'^2 = A^2/Z_m^2.$$

The substitution of (16) so as to eliminate Z_m gives

$$R'^2 + X'^2 = AR'/r. \tag{18}$$

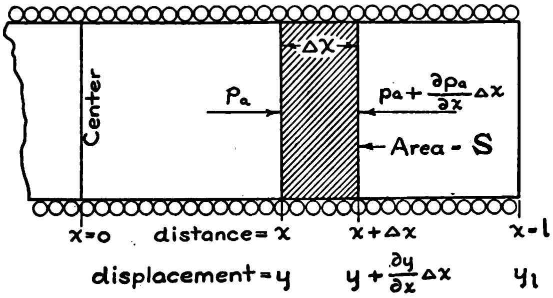

Fig. 13

This is the equation to a circle, passing through $R' = 0$ and $X' = 0$, and having a diameter A/r, and is represented as the left-hand circle of Fig. 11. Each value of ω gives a point of the circle. The circle is thus the locus of X' plotted against R', or, otherwise stated, it is a vectorial plot of z' as ω varies.

Impedance Circle-Diagrams With Diameter Dipping. In the preceding section the diameter of the impedance circle lies along the axis of motional resistance. Experimentally this is

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found to be not the fact. The diameter of the circle usually dips by an angle which we shall call β.

This result can be obtained theoretically if we let one or more of the quantities a, a', a'', L' be a complex quantity independent of ω so that

$$Ae^{-i\beta} = aa'a''L'S/l. \tag{19}$$

This is equivalent to multiplying every vector z' by $e^{-i\beta}$ which gives the whole diagram the form of the right-hand circle of Fig. 11.

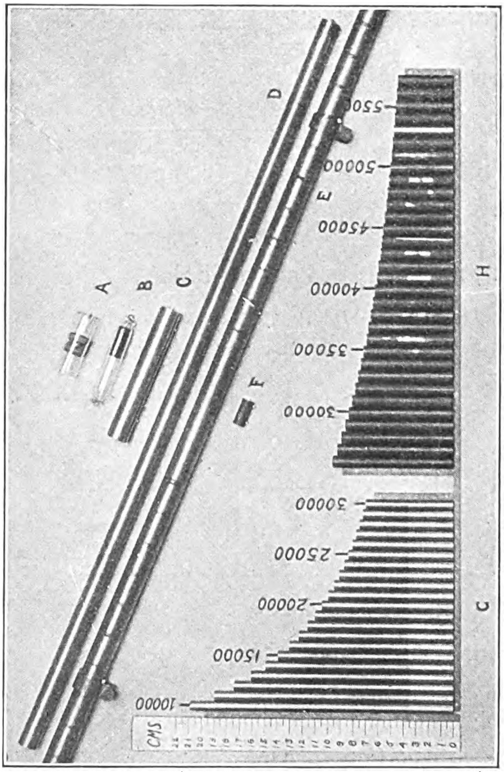

Fig. 14

As to the physical significance of complex values of a, a', a'', or L', which may be inferred from the experimentally determined values of impedance circles, it seems probable that a and a' are the complex quantities and that their complexity is due largely to hysteresis effects and eddy currents in the magnetostrictive

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core as the dip angle largely disappears when the core is a thin nickel tube split longitudinally.

An extensive experimental study of motional impedance in the case of magnetostriction has been made at this laboratory by Mr. K. C. Black[4] at my suggestion, and is contained in his thesis for the Doctorate of Philosophy deposited in manuscript in the Harvard Library. He is publishing a short account of this investigation in a paper following this in the *Proceedings* of the American Academy.

The results are very remarkable in showing that the magnetostrictive system has a very high efficiency as a converter of electric energy into mechanical motion.

Magnetostrictive Sources of Sound. I have made a number of applications of magnetostriction to the production of sound

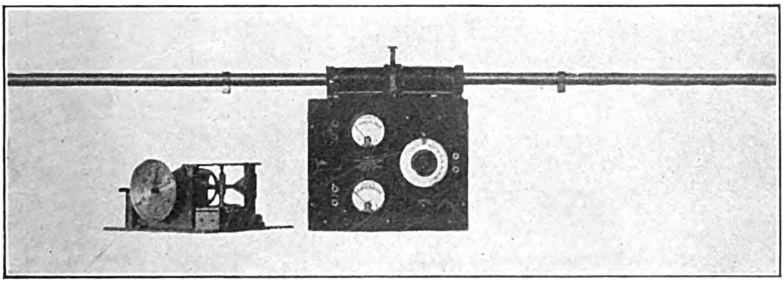

Fig. 15

and ultra sound over a wide range of frequencies. These devices are outside of the scope of the present paper. One application, which is to the measurement of the velocity of sound at various frequencies, is being employed by some of my graduate students in filling in the gaps of the work I previously published[5] in which a piezo-electric crystal oscillator was the radiator of the sound, and the reaction of the reflected sound on the source was the means for indicating nodes and loops of the standing wave system. The frequencies used in those experiments extended from 40,000 to 1,400,000 cycles per second. The use of the magnetostriction vibrators in the same general arrangement of apparatus permits the use of frequencies below the 40,000-cycle

[4] K. C. Black: A Dynamic Study of Magnetostriction, *Proc.* Am. Acad. of Arts and Sci., 63, 49, 1928. These numbers may be purchased separately from the American Academy, 28 Newbury St., Boston, Mass.

[5] *Proc.* Am. Acad., 60, No. 5, 1925.

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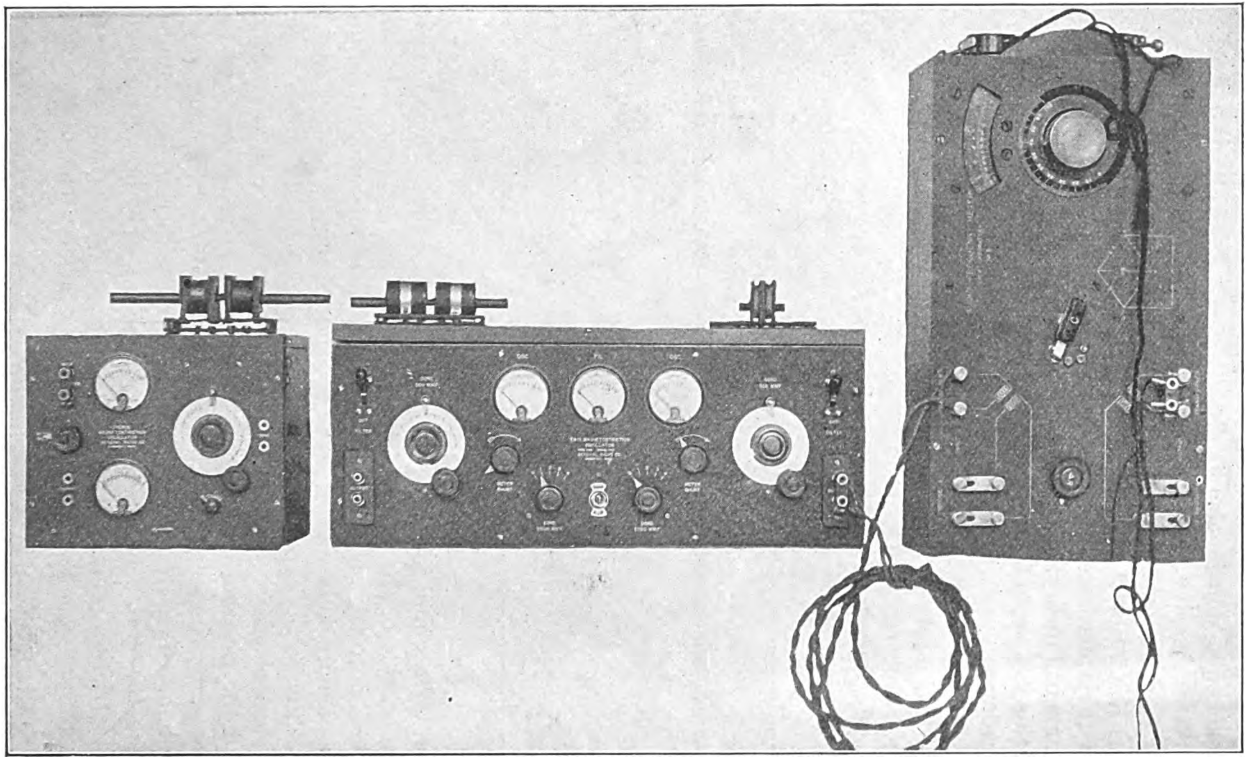

Fig. 16

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value, and also, since rods are easily cut to any predetermined frequency, permits filling in certain significant frequencies that were lacking in the previous investigations. Also various gases are being employed in the experiments.

Another sound-velocity application consists in putting various solid non-magnetostrictive rods on the end of a magnetostrictive driving rod, and by means of frequency measurements determining sound-velocity and elastic properties of non-magnetic and other non-magnetostrictive materials. These new investigations are well under way.

Arrangements for Obtaining High Frequencies. In the production high-frequency stabilization by magnetostriction rods I have used three methods. (*a*) One is by the use of very short cylinders, placed within the plate coil with their axes parallel to the magnetic field of this coil; (*b*) another is by flat strips or sheets of the magnetostrictive material placed between the plate and grid coils; and (*c*) by what I may call *beaded rods.* The latter method is probably the best. An illustrative example, shown in Fig. 12, consists of a rod 1.26 cm in diameter and 9 cm long with grooves turned in it to a depth of half the radius. The grooves are 1 cm wide and separated by 1 cm so that the vibrator is in the form of five cylindrical beads, each 1 cm long held in succession 1 cm apart by cylindrical rods of half the diameter of the beads.

When this beaded rod has been annealed and is placed in a coil-pair having 200 turns on each coil, and so positioned that one bead is in one coil and the next bead is in another coil or else one bead between coils, it shows strong stabilization of frequency 295,480 cycles per second. The material is stainless steel with velocity of 5430 meters per second, so the frequency is that of a linear vibrator of length $543000 \div 2 \times 295480 = 0.918$ cm, which is about 10 per cent less than the computed value of one of the beads standing alone and regarded as a linear vibrator. There are other frequencies of the system, particularly one at 19888 cycles per second.

The theory of such a beaded vibrator has been pretty well worked out but to save space is not here given.

For lack of space also the general vacuum-tube theory is here omitted.

In the accompanying appendix a more detailed theory of sound propagation in the magnetostriction rod is presented.

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Appendix I

Theory of Dynamic Magnetostriction, Including Theory of Propagation of Sound in a Viscous Magnetostrictive Medium. Referring to Fig. 13, let us consider a rod of length $2l$ wound with a solenoid of N turns per cm.

Let S = area of cross section of the rod. Let x = the distance from the center to any section of the rod, and let us consider a volume of length Δx extending from x to $x+\Delta x$.

Let p_a and y = external pressure and displacement at x of the section Δx, and

$p_a+\frac{\partial p_a}{\partial x}\Delta x$ and $y+\frac{\partial y}{\partial x}$ = the corresponding quantities at $x+\Delta x$,

and let us note that the pressure at x and $x+\Delta x$ are both estimated inward upon the section Δx, while the displacements are both in the x-direction.

Before analysing the external pressure p_a into its components, we may equate the force in the x-direction to the mass $\times$ acceleration of the elemental volume Δx, obtaining

$$S\left[p_a-\left\{p_a+\frac{\partial p_a}{\partial x}\Delta x\right\}\right]=S\Delta x\rho\frac{\partial^2 y}{\partial t^2}, \tag{1}$$

in which ρ = density of the material of the volume element.

Performing the operations of (1), dividing by $S\Delta x$ and taking the limit as Δx approaches zero, we obtain

$$-\frac{\partial p_a}{\partial x}=\rho\frac{\partial^2 y}{\partial t^2} \tag{2}$$

Let us now analyse p_a into its constituents; namely, let

$$\begin{aligned} p_a &= p_\epsilon+p_G+p_m, \text{ where} \\ p_\epsilon &= \text{pressure due to elasticity,} \\ p_G &= \text{pressure due to viscosity,} \\ p_m &= \text{pressure due to magnetostriction.} \end{aligned} \tag{3}$$

These various p's are due to the pressures of expansion of an element to the left of that shown in the figure, and are given by the equations

$$p_\epsilon=-\epsilon\left\{\frac{\text{increase of volume}}{\text{volume}}\right\}=-\epsilon\frac{\partial y}{\partial x}, \tag{4}$$

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$$p_G = -G\frac{\partial}{\partial t}\left\{\frac{\text{increase of volume}}{\text{volume}}\right\} = -G\frac{\partial}{\partial t}\left(\frac{\partial y}{\partial x}\right), \tag{5}$$

$$p_m = aB, \text{ in which} \tag{6}$$

ϵ = Young's modulus in dynes per sq. cm

G = a viscosity coefficient in *c.g.s.* units

B = magnetic induction at x, producing p_m

a = a coefficient relating expansive pressure of magnetostriction in dynes per sq. cm. to magnetic induction B.

The coefficients ϵ and G are positive, a is positive for some magnetostrictive materials and negative for others. Also a is not exactly constant but varies with B as well as with the material. We shall, however, need to consider it a constant where B is a small induction added to an already existent constant large induction.

The substitution of (4), (5), and (6) into (2) and (3) gives

$$\rho\frac{\partial^2 y}{\partial t^2} = \epsilon\frac{\partial^2 y}{\partial x^2} + G\frac{\partial\partial^2 y}{\partial t\partial x^2} - a\frac{\partial B}{\partial x} \tag{7}$$

and

$$p_a = -\epsilon\frac{\partial y}{\partial x} - G\frac{\partial\partial y}{\partial t\partial x} + aB. \tag{8}$$

We shall now further analyze B. We shall suppose the existence of a steady magnetic induction B_o, which will polarize the system and determine the size of the coefficient a, but will not contribute to p_m, which we shall suppose to be due to an additional periodic B. This additional B is made up of a quantity due to the variable current i in the windings plus an additional quantity B' called into play by magnetostriction as a result of distortion; that is

$B = B_i + B'$ where

B_i = magnetic induction due to i (9)

B' = magnetic induction due to expansion of the element Δx and is a function of x.

This B' we shall assume proportional to increase of length of the element divided by the original length of the element; that is,

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$$B' = a' \frac{\partial y}{\partial x} \tag{10}$$

In substituting (9) and (10) into (7) and (8) we shall regard

$$\frac{\partial B_i}{\partial x} = 0 \tag{11}$$

The substitution then gives

$$\rho \frac{\partial^2 y}{\partial t^2} = \epsilon' \frac{\partial^2 y}{\partial x^2} + G \frac{\partial \partial^2 y}{\partial t \partial x^2} \tag{12}$$

and

$$p_a = -\epsilon' \frac{\partial y}{\partial x} - G \frac{\partial\, \partial y}{\partial t \partial x} + aB_i \tag{13}$$

where as an abbreviation

$$\epsilon' = \epsilon - aa' \tag{14}$$

Equations (12) and (13) are to be treated as simultaneous.

In solving (12) we shall here restrict the treatment to the case in which the time t enters in y only in a factor of the form $e^{j\omega t}$, where ω is the angular velocity of the current i. This reduces (12) to the form

$$-\rho\omega^2 = (\epsilon' + jG\omega) \frac{\partial^2 y}{\partial x^2} \tag{15}$$

which gives

$$y = e^{j\omega t} \left\{ Ae^{kx} + A'e^{-kx} \right\} \tag{16}$$

with

$$k = +j\omega \sqrt{\frac{\rho}{\epsilon' + jG\omega}} \tag{17}$$

We shall next stipulate that the center of the rod has zero motion; that is, $y = O$ at $x = O$. This gives

$$y = Ae^{j\omega t} \sinh kx \tag{18}$$

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Now to determine A, we introduce the further boundary condition that $p_a = O$ at $x = l$, so (13) gives

$$O = -(\epsilon' + jG\omega)\frac{\partial y}{\partial x} + aB' \quad \text{at} \quad x = l \tag{19}$$

Replacing y in this equation by its value from (18) we obtain

$$Ae^{j\omega t} = \frac{aB'}{k(\epsilon' + jG\omega)\cosh kl}$$

and consequently by (18)

$$y = P \sinh kx \tag{20}$$

where as an abbreviation

$$P = \frac{aB'}{k(\epsilon' + jG\omega)\cosh kl} \tag{21}$$

Let us now turn to a consideration of the electrical system in which there is a variable current i flowing. The variable emf at the terminals of the coil is

$$e = Ri + \frac{\partial}{\partial t}\{\text{flux linkage of } (B_i + B')\} \tag{22}$$

The first two terms of the right-hand side of this equation give e damped; that is, the value of e if the rod were restrained from vibrating. The last term is the contribution of the motion, and may be called the *motional* emf, so

$$e' = \frac{\partial \phi}{\partial t}, \text{ where} \tag{23}$$

e' = motional emf

ϕ = flux linkage of B' with the circuit.

To obtain ϕ, let us note that per unit area of S at any element Δx, distant x from the center, magnetic induction of amount $B' - (B' + \Delta x \partial B'/\partial x)$ leaks out between the turns of wire and these return to the core at distance minus x from the center, and thus link with $2Nx$ turns giving

$$\Delta \phi = -2NSx\frac{\partial B'}{\partial x}\Delta x,$$

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where N = number of turns of wire per cm.

To get the total linkage we must integrate this from $x=O$ to $x=l$, and must add $2NSlB_l'$ to account for leakage from the end, obtaining

$$\phi=2NS\left\{-\int_0^l x\frac{\partial B'}{\partial x}dx+lB_l'\right\} \tag{24}$$

We may first integrate this by parts by putting

$$u=x,\ dv=\frac{\partial B'}{\partial x}dx,\ \text{obtaining}$$

$$\phi=2NSl\int_0^l B'dx$$

We may now complete the integral by replacing B' by its value from (10) and obtain

$$\phi=2a'NSy_l,\ \text{where} \tag{25}$$

$$y_l=\text{value of } y \text{ at } l$$

By (20) and (21) this gives

$$\phi=2a'NSP\sinh kl$$

$$=\frac{2aa'NSB_i}{k(\epsilon'+jG\omega)}\tanh kl$$

Now returning to the motional emf, equation (23), and noting that the time derivative is $j\omega$ times the quantity, and then replacing $j\omega/k$ by its value from (17), we obtain

$$e'=\frac{2aa'NSB_i}{\sqrt{\rho\{\epsilon'+Gj\omega\}}}\tanh kl \tag{26}$$

If now

L = self inductance of the coil with rod damped, we may write Li = flux linkage of Bi, which is $2NSlB_i$, so

$$B_i=\frac{Li}{2NSl} \tag{27}$$

which substituted into (26) gives

$$e'=Qi\tanh kl \tag{28}$$

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In this equation I have introduced as an abbreviation

$$Q = \frac{aa'L}{l\sqrt{\rho\{\epsilon' + jG\omega\}}} \tag{29}$$

Motional Impedance. If we let z' = complex motional impedance of the system, and compare (28) we have

$$z' = Q \tanh kl \tag{30}$$

Equation (30) gives the complex motional impedance of the system; Q and k are defined by (29) and (17), respectively.

To investigate the manner in which z' varies with ω, we may note that while Q involves ω, the term in which the ω appears is comparatively small, and since the whole motional action is confined to a very small range of ω, we may regard Q as a constant.

Let us now write[6]

$$k = \alpha + j\beta, \text{ where} \tag{31}$$

$$\alpha = \omega\sqrt{\frac{\rho}{\epsilon'(1+h^2)}}\, g(h) \tag{32}$$

$$\beta = \omega\sqrt{\frac{\rho}{\epsilon'(1+h^2)}}\, f(h) \tag{33}$$

$$f(h) = \sqrt{\frac{\sqrt{1+h^2}+1}{2}} \tag{34}$$

$$g(h) = \sqrt{\frac{\sqrt{1+h^2}-1}{2}} \tag{35}$$

$$h = G\omega/\epsilon' \tag{36}$$

Introducing the value (31) for k into (39) and breaking z' into its resistance and reactance components, R' and X', we have

$$z' = R' + jX' = Q\frac{\sinh(\alpha + j\beta)l}{\cosh(\alpha + j\beta)l}$$

[6] Regarding the functions $f(h)$ and $g(h)$ see G. W. Pierce: "A Table and Method of Computing Electric Wave Propagation, Transmission Line Phenomena, Optical Refraction, and Inverse Hyperbolic Functions of a Complex Variable", *Proc.* Amer. Acad., 57, No. 7, 1922, where a table of the functions f and g is given for a large range of values of h.

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which after expansion and rationalization becomes

$$R'+jX'=Q\,\frac{\sinh \alpha l \cosh \alpha l + j \sin \beta l \cos \beta l}{\sinh^2 \alpha l + \cos^2 \beta l}$$

From this we may obtain

$$R'^2+X'^2=2R_oR', \text{ where} \tag{37}$$

$$2R_o=Q\,\frac{\sinh^2 \alpha l+\sin^2 \beta l}{\sinh \alpha l \cosh \alpha l} \tag{38}$$

Equation (37) is in the form of the equation of a circle of radius R_o. Equation (38) shows, however, that R_o involves the variable parameter ω so (37) is not a true circle. Nevertheless, in the case of the magnetostrictive system, the resonance is so sharp that the whole phenomenon of motional reactance is exhibited within a frequency range of a fraction of one per cent of the resonant frequency; so the variation of the coefficient R_o is inappreciable to the limit of the errors of the experimental measurements that Dr. Black has made.[7]

[7] K. C. Black, "A Dynamic Study of Magnetostriction," *Proc.* Amer. Acad., **63**, 49; April, 1928.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

THE IMPORTANCE OF RADIOTELEGRAPHY IN SCIENCE*

By

JONATHAN ZENNECK

(President, Institute of Technology, Munich, Germany)

WHEN discussing today what science owes to radiotelegraphy I feel that I must first apologize for having selected so general a subject. While most papers read before this Institute refer to special problems of radiotelegraphy proper, in view of the somewhat extraordinary character of this meeting I hope that an exception to this rule may be permitted, since I am not a professor of radiotelegraphy but of general experimental physics.

There is no doubt that the splendid development of radiotelegraphy itself would be the most attractive theme for a paper, so attractive indeed that it has almost become the standardized subject of addresses to be delivered before radio meetings. Having for a long time been a member of your Committee on Standardization I certainly appreciate standardization, but when it comes to papers, I prefer individual production.

After touching upon the development of radiotelegraphy let me say a few words on the "romantic period in the radio field," as Dr. Alfred N. Goldsmith put it in his Presidential address, and let me refer to it in a somewhat romantic manner. In 1900 we were very proud when, in experiments on the North Sea with a Braun transmitter, we had succeeded in establishing "wireless communication" between the isle of Heligoland and the coast, a distance of some thirty miles. Please do not ask questions as to what "wireless communication" meant at that time. Just believe me that what we had was no worse nor better than anything then given the grand name of wireless communication. Our aim was to get 100 miles. Very soon this was attained by us and by others, and as distances at which wireless messages could be transmitted gradually increased, the goal was pushed further and further, and finally the ideal of enthusiastic wireless men became πR, where π means the cyclic number and R the radius of the earth. In other words, they dreamt of a

* Dewey decimal classification: R010. Original manuscript received by the Institute, August 28, 1928. Delivered before New York meeting of the Institute, September 5, 1928.

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wireless communication with the Antipodes, just as some enthusiasts are dreaming now of rocket airship traffic with the Moon. Then came the time when radio communication with the Antipodes was established by means of short waves with much less power than anybody would have thought of. This result was greatly deplored by all true radio enthusiasts because no further record seemed to be possible, since for mathematical and geophysical reasons, respectively, neither π nor R could be increased. But wireless telegraphy went beyond this limit which nature seemed to have put to any further extension. All of you know that it has become possible to receive radio waves which have travelled at least twice around the earth. All this sounds like a romance, but it is not. With what has been achieved since Hertz's laboratory experiments and Marconi's invention up to the present time, radiotelegraphy means a development unrivaled in the history of natural and technical science.

But I do not want to dwell upon that any longer. What I want to do is to illustrate by a number of examples the influence which this development of radiotelegraphy has had on science and, in the first place, on physics, the mother of wireless telegraphy.

I.

I have to begin with that domain of physics in which wireless telegraphy was raised and which naturally first felt the influence of the new art,—I mean the theory of oscillations.

The old equation of Kirchhoff and Lord Kelvin:

$$L\frac{di}{dt}+Ri+\frac{1}{C}\int i\cdot dt=e$$

determining the free or forced oscillations of a circuit respectively depending on whether or not the impressed emf e is equal to zero, still holds. But owing to the experience gained in radiotelegraphy we know that this equation covers possibilities never thought of at the time when it was developed. At that time it would have been almost absurd to mean by L, R, and C anything but constants of the circuit independent of the current in it. Thus for free oscillations the result was $i=Ie^{-\delta t}\sin(\omega t+\alpha')$ i.e., an oscillation having the cyclic frequency $\omega=1/\sqrt{LC}$ and an amplitude decreasing with time according to an exponential law. For dozens of years this was considered the only solution to the above equation.

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But very soon after the beginning of radiotelegraphy, at the time when the transmitters with their crashing and blinding sparks appeared to be more nearly lightning factories than technical apparatus, there arose some doubt as to whether the previously mentioned solution held for circuits including a spark gap. It was at least not self-evident that the energy absorbed by a spark during a time element dt should be determined by the

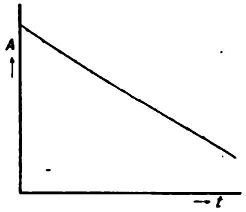

Fig. 1 —Decay of Amplitude in Circuit Including a Spark.

same law as that in an ohmic resistance and hence could be expressed by a term Ri^2dt, where R is a constant. By means of a Braun cathode-ray tube the author made careful measurements with a circuit containing a spark gap and very little ohmic resistance, and plotted the amplitude of its oscillations against time. I found a curve like that in Fig. 1, which is copied from my original paper[1] and which shows that the decay of amplitude

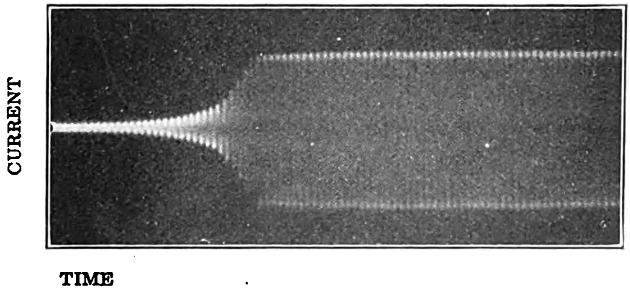

Fig. 2—Increase of Current at Closing an Arc Generator.

follows a linear rather than an exponential law, which means that in a circuit containing a spark the term R cannot be regarded as a constant.

Interesting as this may have been from a physical point of view, practically it does not make much difference whether the amplitude decreases according to a linear or an exponential

[1] J. Zenneck, *Ann. Phys.*, 13, 822, 1904.

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law. When, however, the arc and tube generator came into view, it meant radically new and hitherto unthought-of possibilities. If, for instance, the circuit of an arc generator is closed we get in the circuit an oscillatory current as shown in the oscillogram of Fig. 2 (which the author took a long time ago with a Braun tube[2]); and the corresponding curves for a tube generator are very similar. From such curves it is seen that the amplitude first increases and finally reaches a constant value. If we maintain, at least formally, our equation above, the resistance term R has first a negative value, which gradually decreases to zero. If some thirty years ago someone had proposed to discuss the above equation under the assumption of a negative resistance R, it may be that a mathematician would have been willing to do so, because it is irrelevant to him whether a term has a posi-

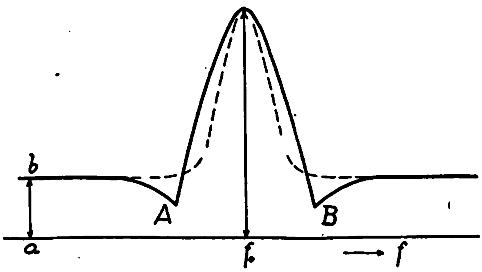

Fig. 3—Resonance Curve of a Tube Generator when Impressed by an emf of Constant Amplitude but Variable Frequency f.

tive or a negative sign. But I am sure that a physicist would have indignantly refused to consider a negative resistance which he would have looked upon as a physical absurdity. In the meantime we have learned that in an oscillatory circuit, when properly connected to an arc or tube, the energy delivered may exceed that absorbed in the circuit and that we might well bring such a circuit within the scope of our equation by attributing to it a negative resistance the value of which decreases with increasing amplitude of the current.

Moreover, if in such a circuit with self-sustained oscillations of frequency say f_0 an emf is applied, the frequency f of which is being gradually varied, we expect to get in the system two oscillations, one being the self-sustained free oscillation of constant frequency f_0 and another the forced oscillation of the variable frequency f. We further expect that the effective value of

[2] J. Zenneck, *Ann. Phys.*, **43**, 481, 1914.

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the current will be the resultant of the two oscillations and therefore that the resonance curve will have the shape of the dotted curve in Fig. 3, where *ab* corresponds to the effective value of the sustained oscillations of frequency f_0. What we actually get[3] is the solid curve of Fig. 3. When we examine the frequency of the oscillations produced we find that between the points *A* and *B* there are no free oscillations of frequency f_0, but only forced oscillations of the variable frequency *f*. This phenomenon is well-known in oscillating audion reception. As Dr. B. van der Pol[4] has shown, it may be interpreted as another result of a resistance the sign and value of which depend upon the amplitude of the current.

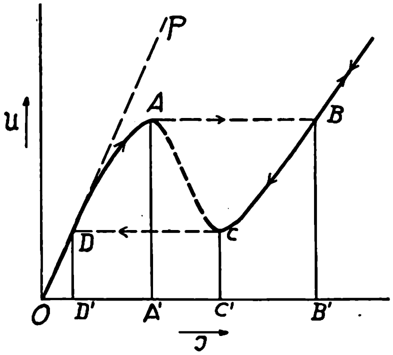

Fig. 4—Voltage-Current Characteristic of a Circuit Containing a Closed Iron Core.

In addition to systems with variable resistance we have to deal with circuits of variable inductance such as those containing an iron core as is used in magnetic frequency changers. Now, in a circuit with constant resistance and inductance the voltage-current characteristic of forced oscillations is determined by

$$u = i\sqrt{R^2 + j\left(\omega L - \frac{1}{\omega C}\right)} \qquad (8)$$

giving a straight line such as the dotted line *OP* in Fig. 4, when the impressed emf *u* is plotted against the current. Instead of that, the characteristic of a circuit containing an iron-core coil has the form of the solid curve of Fig. 4. It exhibits a materially

[3] Vincent, *Proc.* Phys. Soc. London, **32,** 84, 1920; F. Rossmann and J. Zenneck, *Jahrb. d. Draht. Tel.,* **24,** 47, 1924.

[4] B. van der Pol, Jr., *Tijdschr. Nederl. Radiogenootschap,* **2,** 56, 1924; *Phil. Mag.* (7) **3,** 65, 1927.

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new effect[5] which is easily seen when we gradually increase the voltage. Up to the value AA' the current also gradually increases, but when the voltage exceeds this value by a small amount, the current suddenly jumps from the value OA' to the much higher value OB'. This effect which we call *kipperscheinung* (unbalance or instability effect) may be very dangerous, since the heavy current and the correspondingly high voltages at the inductance and condensers may be damaging to the plant. In 1915 when making our first experiments with the machine transmitter at Sayville, in which magnetic frequency changers were used, we got this effect. I well remember how frightened we were when our ammeter suddenly jumped to about double the normal value after we had increased the voltage by an inappreciable amount. We also saw that when the value of the

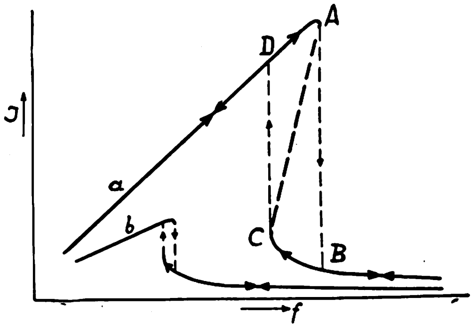

Fig. 5—Resonance Curves of a Circuit Containing a Closed Iron Core when Impressed by an emf of Constant Amplitude but Variable Frequency *f*.

impressed voltage was kept constant while its frequency was gradually varied, we got resonance curves of the kind drawn in Fig. 5 for two different voltages. In consequence of this experience and in order to avoid trouble caused by our lack of knowledge of the working conditions of such a transmitter, I tried to treat the matter theoretically and on September 1st, 1915, almost exactly thirteen years ago, I read a paper before this Institute on "A Contribution to the Theory of Magnetic Frequency Changers" in which I reported the results of my theoretical investigation.[6]

[5] O. Martienssen, *Phys. Zeits.*, **11**, 448, 1910. H. Schunck and J. Zenneck, *Jahrb. d. Draht. Tel.*, **23**, 63, 1924.

[6] J. Zenneck, Proc. I. R. E., **8**, 463, 1920. Paper read before the Institute, September 1, 1915.

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In a loaded frequency changer, we are of course concerned with coupled circuits. The situation with coupled circuits also has changed materially since pre-radio times. The very fundament of previous coupling theory has been touched upon. Take for instance inductive coupling such as shown in the upper part of Fig. 6. It consists of an

$$\text{emf} = -L_{12} \cdot \frac{di_1}{dt}$$

impressed on the secondary from the primary circuit and of a reactive

$$\text{emf} = -L_{21} \cdot \frac{di_2}{dt}$$

induced in the primary. Now, in every textbook of physics you find as a basic relation

$$L_{12} = L_{21}$$

Generally speaking, this means that whenever conditions are such as to produce a strong effect by the primary on the secondary circuit, it necessarily implies a strong reaction of the secondary on the primary. This, of course, does not hold today. As far as I am aware Mr. Armstrong[7] was the first to call attention to the fact that by coupling two circuits through an amplifier

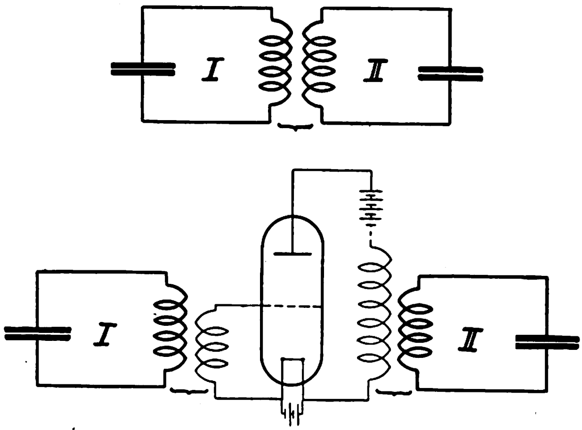

Fig. 6—Upper Part: inductive coupling of two circuits. Lower Part: inductive coupling of two circuits through an amplifier tube.

[7] E. H. Armstrong, Proc. I. R. E., 5, 145, 1917.

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tube, as shown in the lower part of Fig. 6, it is readily possible to have the secondary circuit strongly influenced by the primary without at the same time experiencing any appreciable reaction from the secondary on the primary.

Furthermore it is a well-known result of the theory of tuned coupled circuits, that when the coupling is tight, in the primary as well as the secondary circuit, the same two coupling oscillations appear. Since the frequencies of these two oscillations are different, the time curves of the primary and secondary currents

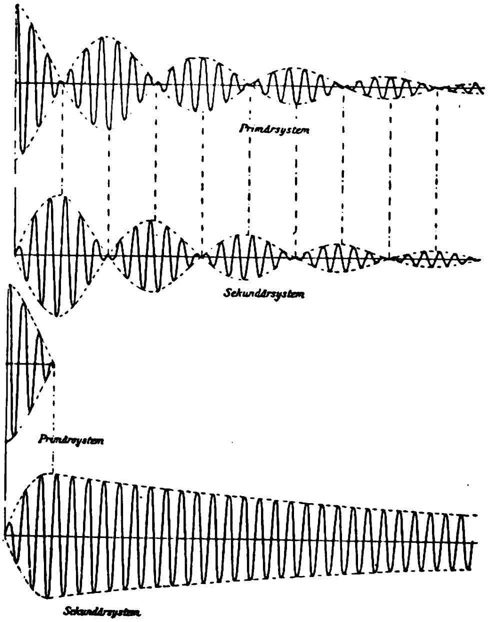

Fig. 7—Upper Curves: damped oscillations in two tightly coupled circuits. Lower Curves: damped oscillations in two tightly coupled circuits, the primary including a quenching gap.

are of the forms of the first and second curves respectively in Fig. 7. We have known for a long time that we may get a totally different effect, as represented by the third and fourth curves in Fig. 7, if in the primary circuit, instead of an ordinary spark gap a proper quenching gap, such as the mercury vapor lamp of Cooper Hewitt or the extremely short gap of M. Wien, is used. In such a case the function of the primary circuit is almost exclusively to excite, by a kind of impulse, the free oscillations of the secondary circuit.

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Some years ago a new form of impulse excitation came up—the magnetic synchronized impulse excitation—in the secondary circuit of which almost undamped oscillations of a frequency much higher than that of the primary circuit may be produced.

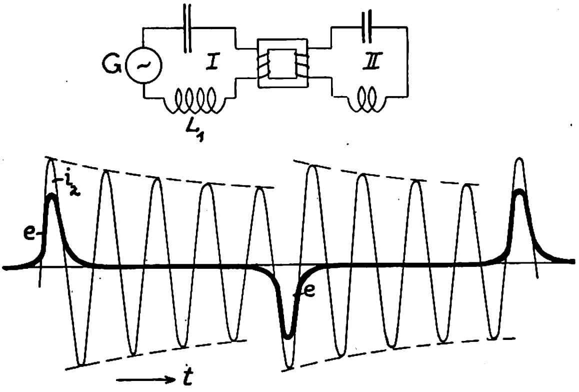

Fig. 8—Magnetic Synchronized Impulse Excitation.

For this purpose in the primary circuit (I, Fig. 8) containing the alternator in addition to a large air-core inductance L_1, the primary of an iron-core transformer is inserted, the iron of which becomes highly saturated by the current used. In this case the

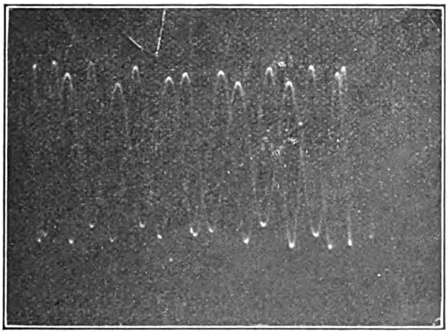

Fig. 9—Magnetic Synchronized Impulse Excitation.
Secondary current plotted against primary current.

emf induced in the secondary (II, Fig. 8) circuit is of the form of the heavy curve *e* of Fig. 8. If, then, the secondary circuit is tuned to a high odd multiple of the primary frequency, its oscillations as shown by curve i_2 in Fig. 8 receive a new impulse twice per period of the primary circuit, and in the meantime between two

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impulses are falling off according to the decrement of the circuit. If this decrement is small or if special means are employed to offset the effect of the natural decrement, almost undamped oscillations in the secondary circuit are produced. This is best proved by the photograph of Fig. 9, a Lissajou figure taken on such a device with a Braun tube, where the secondary current of 15 times the frequency of the primary is plotted against the primary current. In this photograph the damping of the secondary oscillations is noticeable but very small. Incidentally, this photograph with its well-defined clean-cut curve may show you that we have developed the technique of the Braun cathode-ray tube fairly well.

Since the output of the device just mentioned is a current of much higher frequency than that delivered by the alternator, the device may be and generally is considered one for frequency

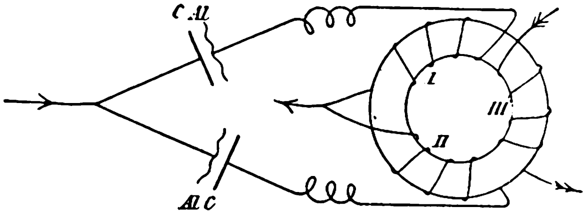

Fig. 10.—The Author's Device for Frequency Doubling.

multiplication. Multiplication of frequency is another field which radiotelegraphy has contributed to the province of oscillations. As far as I am aware, I was the first to disclose the idea of frequency multiplication by static frequency changers. In one of the weekly meetings in the physical laboratory of Professor Braun the advantages of undamped waves for radiotelegraphy had been discussed. There was general agreement that at the time it seemed hopeless to think of generating such oscillations of sufficiently high frequency by an alternator. After this meeting it occurred to me that it might be possible to multiply the frequency of a given alternating current and in a paper[8] which was published in 1899 under the heading "A Method for the Transformation of Frequency by Means of A Static Transformer," I described the arrangement of Fig. 10 which is taken from the original paper. The primary current was divided into two

[8] J. Zenneck, *Ann. Phys. Chem.*, 69, 858, 1899.

branches, each containing an electrolytic valve (the only electric valves then known to me) and a coil, both coils being wound on the same iron core. I showed by oscillograms taken with a Braun tube that in a third coil wound on the same core a current of double frequency was produced. Of course my arrangement had very poor efficiency, but as was frequently the case in radiotelegraphy, when an idea was once proposed, the problem of its realization was attacked upon a number of different lines. Thus some years later I. Epstein described[9] the well-known arrangement of Fig. 11 for frequency doubling. In this system the magnetic properties of an iron-core coil with direct-current saturation are employed. For a dozen years this method has been

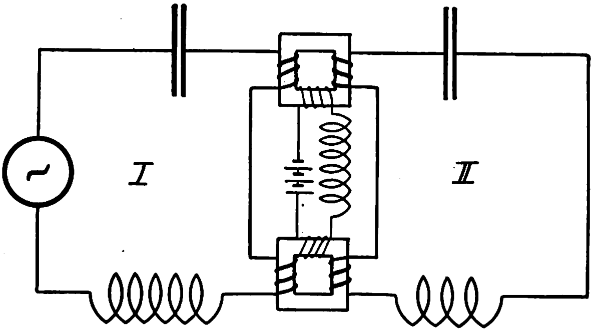

Fig. 11—Epstein's Device for Frequency Doubling.

in constant use for the transatlantic radio service of Nauen and many other commercial stations, including that at Sayville, Long Island from 1915 to 1917.

Now frequency multiplication is but one instance of the interesting frequency situation as generated by the needs and means of radiotelegraphy. That in addition to a frequency multiplication a frequency demultiplication or frequency division should be possible will be obvious to anybody who knows that, mathematically, division is the inversion of multiplication, and who is not aware of how difficult it is sometimes to get the physical equivalent of a mathematical operation. I may refer in this connection to the above-mentioned problem of reversing the sign of a resistance. As a matter of fact present conditions are such that a current of a given frequency may be changed into one of any other frequency. Not only are we able to filter out

[9] J. Epstein, *DRP*, 245; 445 (1903).

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of the oscillations generated by a low-frequency source its high-frequency harmonics, or to produce a low-frequency current by beating two high-frequency oscillations, but we also have means for passing from one audio frequency to another by means of modulating, filtering, and heterodyning radio-frequency currents. I wonder if in the near future music will take advantage of this possibility. For instance by using such a frequency changing device and a loudspeaker a bass singer may, if needed, appear as a tenor or when a flautist should play a number in A flat major which is very uncomfortable on a flute, he might well play it in D major which is much easier and then by a frequency-changing device and loudspeaker transpose it into A flat major.

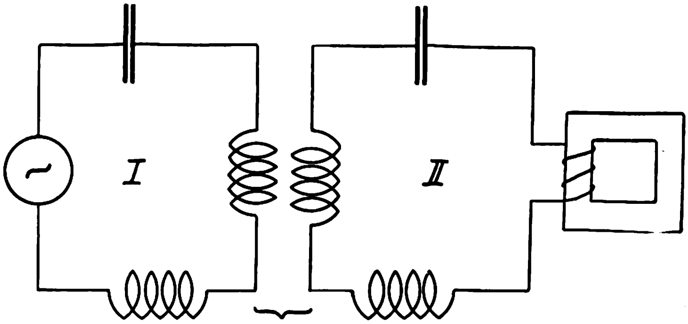

Fig. 12—Circuit Arrangement for Producing Self-modulated Oscillations.

Before leaving frequency changers and coupled circuits I would like to mention a curious effect[10] produced by the harmless-looking arrangement shown in Fig. 12, in which two circuits are tightly coupled, the primary containing the generator, the secondary a coil with closed iron core. Such a device may give currents as shown in the Braun tube oscillograms of Fig. 13. The primary as well as the secondary current seems to have been automatically modulated. This may be of interest for the following reasons: whenever a generator of constant frequency and constant amplitude acts upon a system of circuits it is a general assumption that whatever conditions may be, the currents in these circuits will have constant amplitude, at any rate after the transient phenomena have disappeared. This as-

[10] K. Heegner, *Zeits. f. Phys.*, **29**, 91, 1924. H. Plendl, F. Sammer, and J. Zenneck, *Jahrb. d. Draht. Tel.*, **26**, 104, 1925.

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sumption is expressed by putting the current in any of the circuits in the form

$$i = Ie^{-j\omega t}$$

That there are cases where this almost self-evident assumption does not correspond with the fact is shown by the experiments just mentioned.

These examples may be sufficient to demonstrate how far, compared with the classic theory of oscillations, our knowledge in this field has been extended by the problems and experiences met with in radiotelegraphy.

II.

At least as important as the suggestions and problems which physics was given by radiotelegraphy are the instrumentalities which radiotelegraphy added to the experimental resources of physics.

All of you know how troublesome it was to measure a weak alternating current. While it is a simple matter to measure a direct current of the order of 10^{-9} ampere by a moving coil galvanometer, or even one of the order of 10^{-11} or 10^{-12} ampere with an armored fixed-coil galvanometer having an astatic system, the best we had for the measurement of feeble alternating currents was the vacuum thermo-couple used with a sensitive direct-current galvanometer. Compared, however, with the present means for measuring direct currents this method is extremely insensitive. Nowadays radiotelegraphy has furnished us with means such as good crystal detectors, Fleming valves, De Forest audions, and all those methods known as tube ammeters and voltmeters which have made measuring weak alternating currents and voltages a pleasure.

In addition to rectifiers for feeble currents radiotelegraphy has provided high-vacuum rectifiers for heavy loads such as the Kenotron of the General Electric Company. The modern Roentgen ray methods are made use of extensively to produce direct currents of extremely high voltage needed for operating present day Roentgen ray tubes. These rectifiers have entirely superseded the old synchronous mechanical commutators which were formerly used for rectifying alternating current of high voltage. I am sure that no one working in a physical laboratory will regret the disappearance of this apparatus as the sparks and the

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undesired oscillations they excited in all more or less nearby conductors were about as welcome in a physical laboratory as hail in a corn field. And modern Roentgen ray tubes themselves, with their high vacuum and their heated cathodes such as those developed by Dr. Coolidge, have grown in the same field where the electron tubes of radiotelegraphy were raised.

Coming now to the electron tube as an amplifier I am almost afraid to say anything. So general has become its use in experimental physics that when we talk of a tube we mean an amplifier tube, and when we mention an amplifier, everyone understands that we mean a tube amplifier. When we read a paper on an

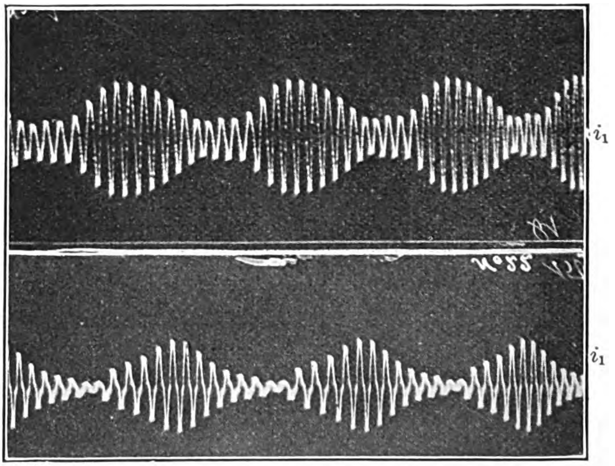

Fig. 13—Primary and Secondary Circuit as Produced by the Device of Fig. 12.

experimental investigation we are greatly astonished if no amplifier tube is used. As a matter of fact with the amplifier tube the experimental situation has become such that almost any effect, feeble as it may be, can be amplified to such a degree as to be measured, oscillographed, made audible or visible. A few instances may illustrate that.

In the photometry of stars, by means of the photoelectric cell, direct currents of the order of magnitude of 10^{-14} ampere have to be dealt with. According to experiments made by H. Rosenberg[11] it is possible without appreciably affecting the proportionality between light intensity and current to amplify these

[11] H. Rosenberg, *Berl. Ber.*, **53**, 716, 1920.

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currents about 100,000 times, so as to measure them by a moving coil galvanometer.

For dosing Roentgen rays it is common practice to measure the current produced by them in an ionization chamber. These currents are of the order of magnitude of 10^{-12} ampere. Commercial apparatus is now being built[12] in which, by a single 4-electrode tube, these currents are amplified and then measured by a direct reading moving coil galvanometer with a sensitivity of 2.6×10^{-7} ampere per scale division. Such apparatus can be handled by any physician while the former electrometric method for measuring these ionization currents was too complicated to be left to a regular doctor.

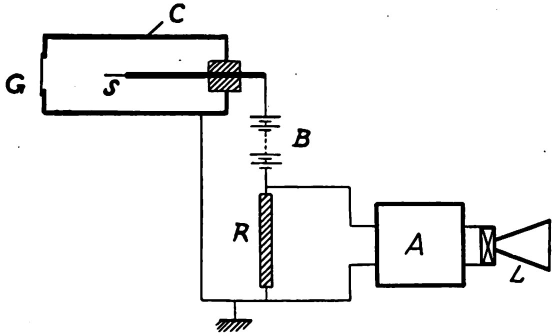

Fig. 14—Geiger Counter.

Another example is the arrangement of Fig. 14 representing the well-known Geiger counter, a metallic cylinder *C* closed at the front by a mica disk *G*, in the interior of the cylinder and insulated from it a metallic point *S* such as a sewing needle, between the point and the cylinder a battery *B* and a high resistance *R*, connected between the grid and filament of the first tube of a multi-stage amplifier *A*, the output of which is fed into a loudspeaker. The voltage of the battery is such that no glow discharge is started between the point and the cylinder. If now a radioactive substance emitting alpha particles is placed near the mica disk, a very loud and sharp crack is produced in the loudspeaker every time an alpha particle is shot through the disk into the interior of the chamber, ionizing the air in it. If Roentgen rays are sent into the cylinder, a great

[12] K. W. Hausser, R. Jaeger, and W. Vahle, *Wissensch. Veröffentl. des Siemens-Konzerns*, **2**, 325, 1922.

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number of cracks are heard corresponding to the separation of electrons from the atoms and their ionization effect. This experiment, which may be made easily in even a large lecture room, represents an interesting acoustic analogue to the well-known experiments of C. T. R. Wilson, by which these effects are made visible.

To take still another example, if the magnetization of a steel bar or steel needle is reversed, the change in magnetization does not take place gradually, but in small steps when, according to Ewing's view, groups of magnetic elements—whatever they may be—tilt over. This effect may be made audible in a large lecture room by an arrangement due to Barkhausen.[13] The steel needles or bars are placed in a coil with a large number of turns; connected to this coil, through a multi-stage amplifier, is a loudspeaker. If now a strong permanent magnet is moved along the coil and steel bars, a roaring is heard in the loudspeaker owing to the discontinuous change in magnetization. I have mentioned these two last devices in order to show how, by means of an amplifier, elementary processes have been made directly accessible, the existence of which could hitherto be inferred only from their integral effect.

The most wonderful achievement in radiotelegraphy in my opinion is the tube generator. In stating that, I do not think of detracting from the merits of those who, following the steps of Nikola Tesla, built alternators for the production of undamped oscillations. On the contrary I feel that these alternators, together with their frequency-controlling devices such as those of Mr. Alexanderson, represent the most ingenious products of the electrical art. I also fully appreciate V. Poulsen's arc generator which with extreme simplicity combines the possibility of very large units. From the very beginning of radiotelegraphy it was a dream of all connected with it, to be able to work with undamped oscillations. In view of the apparent advantages of such oscillations the problem of generating them was tried from all possible lines of attack. I do not think that twenty years ago anyone would have dared to dream of a generator such as the tube generator which, with relatively simple apparatus, would produce undamped oscillations up to a frequency of 30,000,000 per second or more with a constancy of amplitude and frequency unsurpassed by any other generator, and which when required could be modulated very easily by an audio-frequency current.

[13] H. Barkhausen, *Phys. Zeits.*, 20, 401, 1919.

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Of course I am aware that I am not quite correct in saying that the constancy in frequency of the tube generator is unsurpassed by any other generator. As a matter of fact it is surpassed by the tube generator itself when combined with a piezo-electric quartz crystal. I do not desire to dwell upon the practical aspect of this contribution to radiotelegraphy by Professor Cady.[14] What I want to emphasize is its importance from a general point of view. Piezo electricity is one of those physical phenomena which was considered more or less a curiosity even by physicists. Just as in the case of electron emission from hot wires some thirty years ago no physicist, and surely no engineer, would have believed that it would ever be of any practical use. We know now that this phenomenon of piezo electricity has led to an electric oscillator of most valuable properties. It is one of those instances I refer to in warning engineering students in my lectures against the idea that this or that physical phenomenon will never be of practical use and that for this reason they may themselves refrain from being interested in it. Furthermore by the piezo-electric oscillator it has been demonstrated that crystals may be capable of mechanical or elastic vibrations of a frequency of the order of some millions per second and of a decrement as low as 0.0001. I am convinced that no physicist would have believed that this would be possible. I freely confess that if before the piezo-electric oscillator was developed, a gentleman had come to my laboratory and had told me that he had made an invention by means of which it was possible to excite elastic vibrations in a quartz crystal up to a frequency of some millions per second, I would have treated him very politely but would have recommended that he see a doctor.

To come back to the tube generator, I need not tell you that for radio-frequency measurements it has become the only oscillator which is now used. It is especially valuable in connection with Professor R. A. Fessenden's heterodyne principle, not in the original form but in one adapted to this oscillator. As a matter of fact, any minute change in almost any physical property may be detected and measured by this method, since it is practically always possible by this change to influence the frequency of a tube oscillator and thus vary the beat frequency produced by the interaction of this oscillation with one of constant frequency. For instance it is obvious that by this method

[14] W. G. Cady, Proc. I. R. E., 10, 88, 1922.

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the dielectric constant of gases, as dependent upon pressure, may be determined readily by simply putting the condenser of the variable generator into the gas, or by making one plate (*B*, Fig. 15) of the condenser movable; extremely small displacements of it may then be measured. This scheme as represented

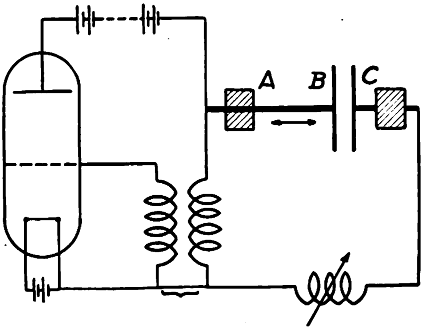

Fig. 15—Device for Measuring Small Displacements.

in Fig. 15 was worked out by R. Whiddington.[15] He succeeded in measuring displacements of the order of 10^{-8} cm, that is, of the dimension of atoms. This means an accuracy far beyond that obtained by light interference methods. That by a similar arrangement a small extension by heat of any solid material,

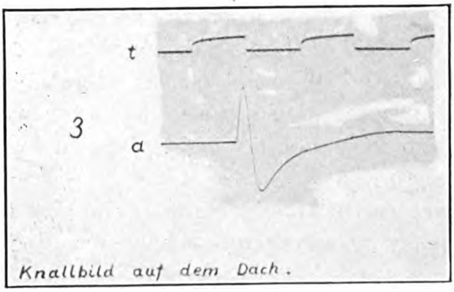

Fig. 16—Curve *t*: time marks of 1/32 sec. Curve *a*: time curve of air pressure in shock test.

that small rotations of parts of seconds of angle,[16] that minute movements of the beam of a highly sensitive balance[17] may be measured, is evident. I only mention it to demonstrate the wide scope of this method.

[15] R. Whiddington, *Phil. Mag.*, 40, 634, 1920.
[16] A. Pflüger, *Phys. Zeits.*, 22, 73, 1921.
[17] F. Kock and G. Schweickert, *Phys. Zeits.*, 23, 123, 1922.

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III.

One of the most striking consequences of radiotelegraphy is the revival of the study of acoustics. For a long time this study had been somewhat discredited, so that a physicist working in this field was suspected of failing to find an important problem. This situation has been entirely changed. On one hand, radiotelegraphy has brought up new acoustical problems such as the proper construction of rooms for broadcasting studios, the design of good microphones and loudspeakers. On the other hand, radiotelegraphy has supplied physical and physiological

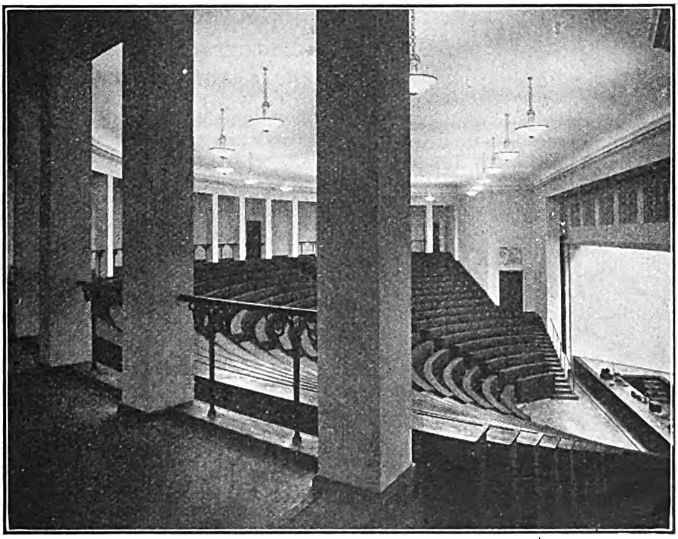

Fig. 17—Lecture Room for Physics at the Institute of Technology in Munich.

acoustics with new experimental means, with its tube generators, its amplifiers, and its high-grade microphones. Thus it became possible to attack successfully problems which a long time ago had been abandoned as practically hopeless, such as the problem of experimentally determining and, if necessary, improving the acoustical properties of a room. To illustrate this I would like to report briefly on experiments[18] made at my suggestion by two of my pupils, Messrs. Scharstein and Schindelin. For the acoustical investigation of rooms we used two methods. In the first method, which we called the shock test, a shot was fired from a

[18] To be published soon in the *Ann. Phys.*

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small 22-calibre pistol and the time curve of the air pressure at some point in the room was registered by means of a Reiss microphone, a multi-stage amplifier with resistance-capacity coupling, and a Siemens oscillograph. The curve received in free air, that is, on the roof of my laboratory, was that reproduced in curve 3*a* of Fig. 16. In the second method, which we called the tone test, an audio-frequency tube generator with a loudspeaker in connection with a rotary interrupter sent out groups of tones at regular intervals, and again the time curve of air pressure as produced by this tone group in the room to be investigated was oscillographed.

I want to show you two results. The first is that of the shock test of my lecture room, a photograph of which is shown in Fig. 17 and an elevation in Fig. 18. As you see from the photo-

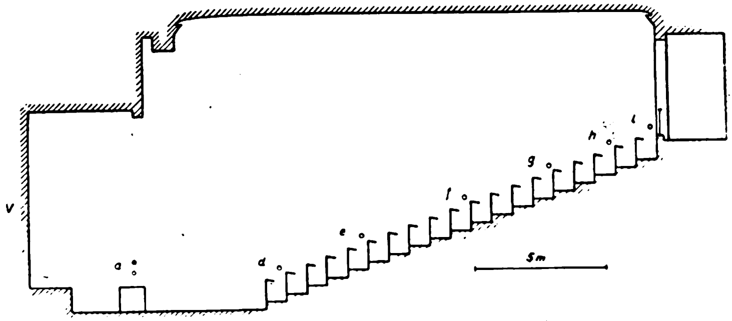

Fig. 18—Elevation of the Lecture Room of Fig. 17.

graph the room resembles an amphitheater of very simple geometrical form. In carrying out the shock test of this room, the pistol was fired at about the geometrical center of the room, i.e., at point *a* of Fig. 18 and close to it was the receiving microphone. The oscillogram as produced in the empty lecture room is that of the second curve in Fig. 19 while the first and third curves represent time marks of 1/32 of a second given by a tuning-fork interrupter. There is to be seen in the oscillogram first the direct shot then the reflections from each row of benches, that from the first row with maximum and those from the others with decreasing intensity, and finally the reflection at the gallery. In the full lecture room as shown by the fourth curve in Fig. 19, the reflections at the benches almost disappear because of the disturbing effect of the students sitting on the benches. There is

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only left that at the front of the first bench and that at the gallery. In view of the accuracy with which all details of such a room are reproduced in the oscillogram, it almost might be called an acoustic photograph although this expression might be seriously objected to from a philolgical point of view. The great advantage of this shock test is that from the time interval between the direct shot and its reflection sometimes in connection with the directive property of the pistol shot, it is possible even in complicated cases to place with great exactness that portion of the room in which a reflection takes place and which may be the source of a disturbing echo in the room. Considering the precision of such measurements, a few weeks ago we had a very curious experience. I received a letter from the University

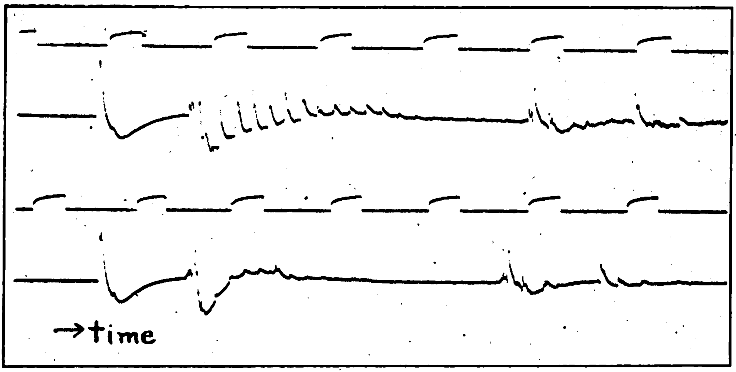

Fig. 19—Oscillogram of Shock Test in the Lecture Room of Fig. 17.

of Freiburg in Baden asking if it was not possible to find out the reason for their auditorium having extremely poor acoustical properties. I sent the above-mentioned engineers there, and they first made a shock test of the auditorium. Their oscillograms showed strong reflections. When, however, they tried to locate the reflecting surface by comparing the time intervals with the plan of the auditorium they met with serious difficulties. They repeated their experiments after having checked their tuning-fork interrupter, but got the same results. Finally they began to doubt the correctness of the plans and measured the main dimensions of the auditorium. They found that the r acoustical measurements had been correct but the plan was inaccurate. Thus, having the correct dimensions, it was easy to find the reflecting surfaces.

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As an example of a tone test I have reproduced in curve 1*a* of Fig. 20 the oscillogram obtained at the studio of the broadcast transmitter in Munich, while curve 1*t* gives the time marks and curve 1*b* the current in the loudspeaker. The oscillogram shows only one weak reflection following the direct sound at a very short interval. The tone groups are clearly separated from each other, corresponding to the relatively good acoustic proporties of the room. In strong contrast thereto is the corresponding oscillogram—(curve 2*a*)—taken in a hall of the Institute of Technology at Munich, having an arched stone ceiling and a stone floor. The oscillogram clearly exhibits the very bad acoustic properties of the room, the tone groups being completely mixed up; with

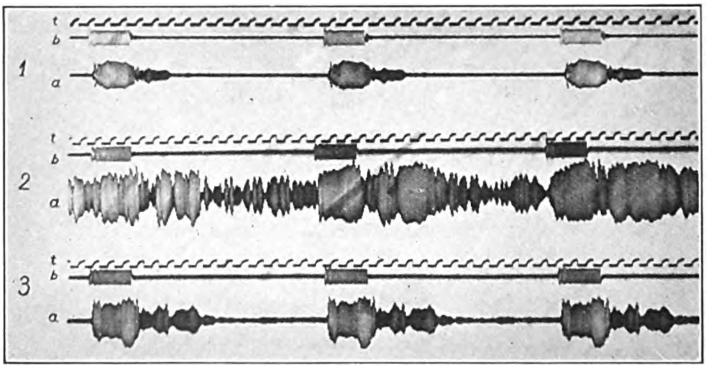

Fig. 20—Oscillograms of Tone Tests. Curve 1: in the studio of the broadcast transmitter in Munich. Curve 2: in an entrance hall of the Institute of Technology in Munich. Curve 3: the same as in curve 2, but the floor having been covered with a sound absorbing material.

the ear tone groups cannot be distinguished at all. Conditions were still worse when the frequency of the tone used coincided with that of a free oscillation of the hall. If now the acoustical conditions were examined by a shock test, it was found that the path of the sound rays was approximately that shown in the lower part of Fig. 21, the loudspeaker being at the point of the shaded circle. As shown by this drawing the sound is first reflected at the arched ceiling, concentrated near the floor and then reflected by it, then reflected again at the ceiling and so on. This goes on for many seonds owing to the absorption by the ceiling as well as that by the floor being very weak. Now it was to be expected that this almost infinite series could be cut off after the first term if the sound could be absorbed at its first arrival

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from the ceiling. We did this by covering the floor with *molton*, which is a thick fabric. The result will be seen from curve 3*a* in Fig. 20. This shows that by simply covering the floor the former extremely bad acoustic properties of the hall were improved to such a degree that they were not much inferior to those of the broadcast studio represented by curve 1*a*. Interesting in this connection is the fact that by covering the floor the bad acoustic in-

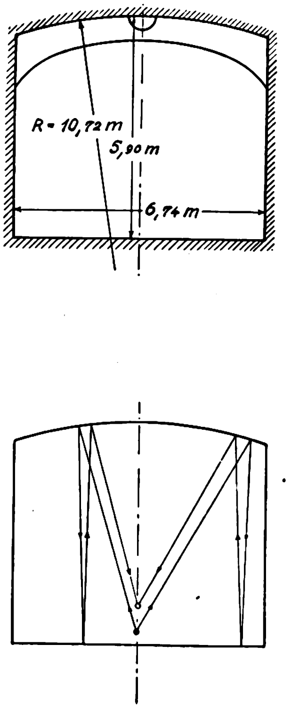

Fig. 21—Elevation of the Hall (to which curves 2 and 3 of Fig. 20 refer) and Path of Sound Rays.

fluence of the ceiling has been eliminated, aside from the first reflection thereon, which however is relatively harmless, since it follows the direct sound at a very short interval. Still more interesting is the following observation made in the course of these experiments. When the source of sound is located at the side of the hall, the sound rays by the first reflection from the arched ceiling are concentrated on a relatively small part of the floor. If, then, this part only is covered by a sound absorbing material, practically the same result is obtained as if the whole floor had

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been covered. In our experiments a piece of *molton* four square meters in size was sufficient. It was very interesting to hear the sound of the interrupted loudspeaker in the hall suddenly change when this piece of *molton* was alternately put down and taken away. There are many instances such as lecture rooms or churches, where the speaker always stands at the same place. From the experience just mentioned it is to be concluded that in such cases it might be possible to improve the acoustic properties of the room by very simple means, of course for only one definite position of the speaker.

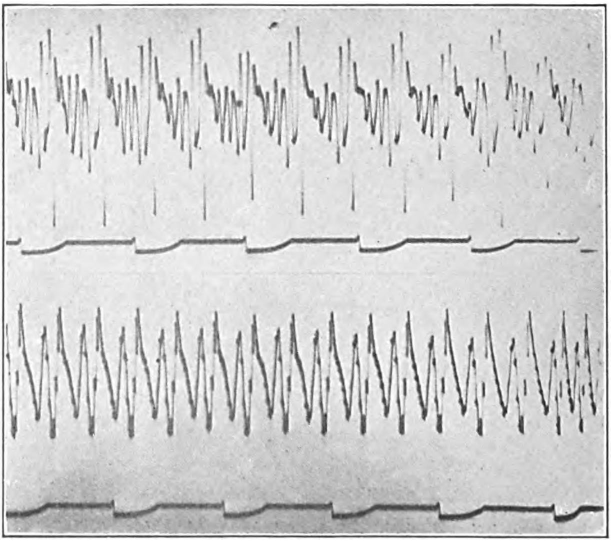

Fig. 22—Time Curves of the Vowels *e* and *a*.

As further evidence of how far the experimental methods furnished by radiotelegraphy came into use in acoustics I want to quote the experiments recently made with oscillations of ultrasonic frequency of the order of some 100,000 per second. How closely modern acoustic methods are coupled with radiotelegraphy can be best proven from the fact that it is now common practice to publish papers on acoustics in journals of radiotelegraphy such as the PROCEEDINGS of the Institute.

It is also but natural that physiological acoustics, which of course is closely related to physical acoustics, enjoys the experi-

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mental means afforded by radiotelegraphy.[19] Let me show just one instance. In Fig. 22 there are reproduced the time curves of the vowels *a* (as in "are") and *e* (as in "Ellen") which were taken in my laboratory with a Reiss microphone, a resistance-capacity coupled multi-stage amplifier, and a Siemens oscillograph. The progress obtained by these modern experimental means cannot be better demonstrated than by comparing those curves showing the finest details of sound with the corresponding curves as formerly taken with the so-called manometric boxes and flames.

It is but recently that medical science is beginning to make use of the apparatus developed in radiotelegraphy and in modern acoustics. To illustrate this I want to show you, from a paper by H. Trendelenburg,[20] two records of heart sounds taken with a

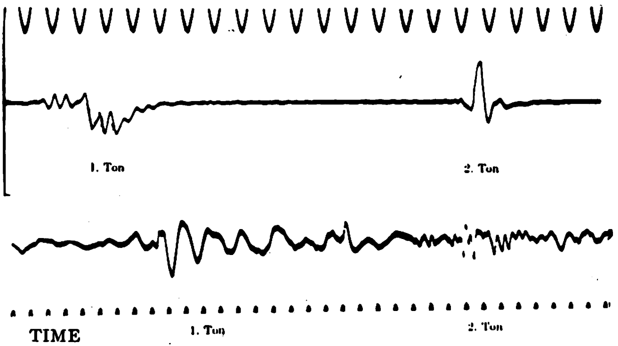

Fig. 23—Time Curves of Heart Sounds.

similar arrangement to that above mentioned except that the microphone used was the excellent high-frequency condenser microphone constructed by H. Riegger of the Siemens and Halske Co. The second curve of Fig. 23 shows the time curve of the heart sound of a healthy man of 31 years while the third curve of the same figure represents the corresponding record obtained with a woman of 70 years of age suffering from a serious heart disease. There is certainly no knowledge of medicine needed to realize that such curves afford a much more reliable basis for diagnosis than mere listening to the sound as was hitherto general practice.

[19] F. Trendelenburg, *Jahrb. d. Draht Tel.*, **28**, 54, 84, 1926.

[20] F. Trendelenburg, *Wissenschaftl. Veröffentlichungen aus dem Siemens-Konzern*, **6**, 184, 1928.

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It would be an attractive proposition to discuss now one science which recently came into close contact with radiotelegraphy—the physics of the atmosphere. There can be no doubt that the recent investigations into the propagation of radio waves are of extreme interest for radiotelegraphy itself. But with regard to the physics of the atmosphere it is hardly possible to draw therefrom any conclusions which go beyond a rather rough estimation of the electron density in different heights. Similar is the case with the influence on the propagation of radio waves of solar activity as discussed in the well-known papers by Dr. L. W. Austin and G. W. Pickard. It seems to me that we are only at the beginning of an interesting development which, however, is encountering special difficulties since it lacks the main requirement of experiments, the modification of conditions at will. It would seem premature to go into any details today.

I have tried to illustrate by a few examples what science in the most general sense of the word owes to radiotelegraphy. The importance of a scientific discovery or of a technical invention may be judged from the practical results obtained, or from the impetus it has given to the development of science and to the enlargement of general human knowledge. From either point of view it seems to me that there is nothing to be compared with radiotelegraphy, and this country and the members of this Institute may be proud of the original ideas they have produced and of what they have done for the scientific and practical development of this art.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

AN AUXILIARY FREQUENCY CONTROL FOR R. F. OSCILLATORS*

By
G. F. Lampkin

Summary—*A method of varying the frequency of an oscillator in small known amounts is described. The control operates on the normally fixed element in the oscillating circuit. The vernier calibration is readily made and maintained.*

DETERMINATION of the overall characteristics of a receiving set requires that some sort of a local modulated oscillator be used to simulate a transmitting station. Determination of the particular characteristic of selectivity necessitates that the oscillator frequency be capable of variation in small steps about any given frequency. To obtain transmission-frequency characteristics of single circuits, band-pass filters, coupled circuits, and so on, an auxiliary control of the

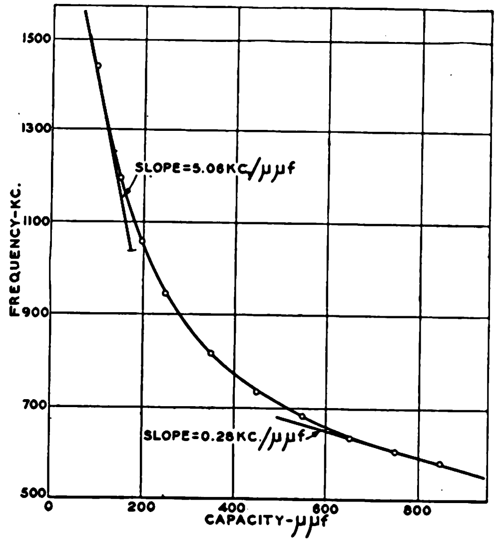

Fig. 1—Frequency-Capacity Relation for Oscillating Circuit.

same sort is needed. It is the purpose of this paper to describe a method for varying the frequency of an oscillator in small, continuous, and known amounts about any frequency in the oscillator range.

* Dewey decimal classification: R351. Original manuscript received by the Institute, September 22, 1928.

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The usual oscillator employs a variable condenser as a means of covering the frequency range. A small semi-circular plate condenser can be used in parallel as a vernier frequency control. The calibrations of frequency change vs. dial setting will be straight lines for any given setting of the main control. However, the value of cycle change per auxiliary-dial division will vary widely over the oscillator range. This is evident from the measured frequency-capacity relation, Fig. 1, for a typical oscillator.

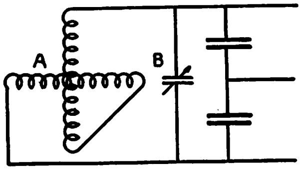

Fig. 2a—Oscillator Circuit with Capacitive Auxiliary Frequency Control.

At 600 kc a small auxiliary condenser would produce a change of 281 cycles per $\mu\mu$f. At 1400 kc the change would be 18 times as great, or 5060 cycles per $\mu\mu$f. This extreme variation in the calibration constant is not only inconvenient in itself, but it lowers the precision of control at higher frequencies.

A remedy can be had in a circuit such as Fig. 2a, or vice versa, 2b. Element *A* in each circuit represents the main tuning control and element *B* the auxiliary frequency control. The auxiliary

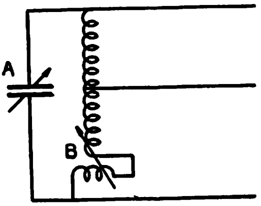

Fig. 2b—Oscillator Circuit with Inductive Auxiliary Frequency Control.

operates on the fixed element of the oscillating circuit. The change in total capacity, or total inductance, due to the auxiliary is then nearly constant over the oscillator frequency range.

Fig. 3 shows the calibration curves that were obtained on an oscillator of the type in Fig. 2a. A variable inductor of 38 to 370 μh was the main tuning control. The double condenser capacity was 320–320 $\mu\mu$f; the maximum capacity of the auxiliary was 8.6 $\mu\mu$f. The calibrations obtained are straight lines; the ratio of the

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slopes at 1400 kc and 625 kc is approximately 2. These auxiliary calibrations were determined by picking up and heterodyning in a detector a small fraction of the oscillator output. The frequency of the beat was measured by sound comparison with a calibrated audio oscillator.

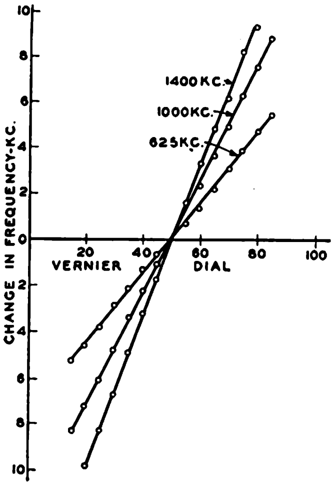

Fig. 3—Calibration of Auxiliary Frequency Control. Slope for 1400 kc = 326 cycles/division; 1000 kc = 249 cycles/division; 625 kc = 154 cycles/division.

Since the change-in-frequency calibrations are straight lines, a value of cycles change per auxiliary-dial division can be determined for each carrier frequency, and the resulting curve plotted. This curve for the particular oscillator is given in Fig. 4. Thus at

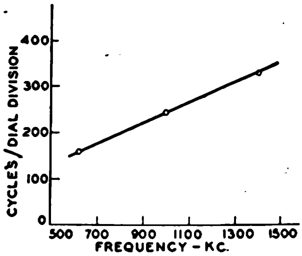

Fig. 4—Calibration of Auxiliary Frequency Control.

any given oscillator frequency the change in frequency due to the auxiliary is known. If the auxiliary control were made a small variometer, and the main control a straight-line-frequency condenser, then the calibrations of both the main and auxiliary frequency controls would be linear.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

A METHOD OF TREATING RESISTANCE STABILIZED RADIO-FREQUENCY AMPLIFYING CIRCUITS*

By
B. L. Snavely and J. S. Webb
(Department of Physics, Lehigh University, Bethlehem, Pennsylvania)

Summary.—*An equation is developed showing the relation between the circuit constants and the critical resistance in a resistance stabilized amplifier having a tuned grid circuit and a pure inductance plate load. In the derivation of this equation it has been found necessary to assume the grid to filament capacity to be zero. Equations are given which may serve as a basis for approximation formulas which will take this capacity into account.*

The experimental results suggest a convenient method of measuring very small capacities, and demonstrate the possibility of controlling regeneration, in this type of circuit, by means of a small condenser between the grid and filament.

Introduction

Recently there has been considerable use of tuned radio-frequency amplifying circuits employing a resistance in the grid circuit for the purpose of preventing oscillation. The type of circuit referred to is shown in Fig. 1. The resistance R has been called the suppressing resistance or the oscillation suppressor, and the method has been termed resistance stabilization or resistance neutralization.

Steady oscillation cannot occur if the value of R exceeds the value of the negative input resistance of the tube. The value of R for which oscillations are just maintained when once started is here termed the *critical value.*

Since the input capacity of the tube is generally not negligible in comparison with C, any increase of R beyond the critical value causes an effective increase in the resistance of the tuned circuit LC and hence results in a decrease in the sharpness of tuning and efficiency of the amplifier. It therefore becomes desirable to have some means of calculating the critical value of R, which we shall call R_c.

Circuit Characteristics

In order that a general expression for R_c might be arrived at, it has been found necessary to make the following assumptions regarding the circuit:

* Dewey decimal classification: R132. Original manuscript received by the Institute, October 6, 1928.

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(a) No current flows between the grid and the filament in the tube.

(b) The plate circuit load is a pure inductance.

(c) The plate to filament capacity is negligible.

(d) The grid to filament capacity is zero.

Assumption (a) is generally fulfilled in most amplifiers.

Assumptions (b), (c), and (d) necessarily limit the direct practical application of the results. Ballantine has shown that the input characteristics of a vacuum tube having an inductance load are not appreciably affected by the resistance of the load provided the resistance does not exceed several ohms.[1] Hence our results may be extended with fair accuracy to the case in which the plate load consists of a pure inductance in series with a pure resistance.

The grid to filament capacity influences the value of the critical resistance quite markedly; hence the results here obtained must be corrected for this capacity.

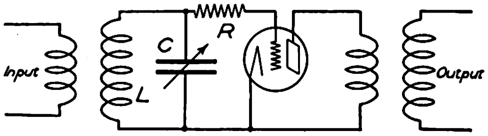

Fig. 1

On the basis of the above assumptions, the circuit analytically equivalent to that of Fig. 1 is shown in Fig. 2. In the latter figure we have:

R_p = plate impedance of the tube.

C_{pg} = grid to plate capacity in the tube.

e_p = a fictitious emf set up within the tube due to the instantaneous grid potential e_g.

Determination of the Critical Resistance

The currents in the circuit will be governed by the following equations:

$$L_1 \frac{d}{dt} i_1 + R_1 i_1 + \frac{1}{C_1} \int (i_1 - i_2) dt = 0 \qquad (1)$$

$$\frac{1}{C_1} \int (i_2 - i_1) dt + R i_2 + \frac{1}{C_{pg}} \int i_2 dt + R_p (i_2 - i_3) = e_p \qquad (2)$$

[1] *Phys. Rev.*, **15**, 409, 1920.

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$$R_p(i_3-i_2)+L_p\frac{d}{dt}i_3=-e_p \tag{3}$$

$$-e_g=Ri_2+\frac{1}{C_1}\int(i_2-i_1)dt \tag{4}$$

$$e_p=\mu e_g \tag{5}$$

These, upon differentiation and elimination of e_g and e_p yield:

$$\left(L_1\frac{d^2}{dt^2}+R_1\frac{d}{dt}+\frac{1}{C_1}\right)i_1-\frac{1}{C_1}i_2=0 \tag{6}$$

$$-\left(\frac{1+\mu}{C_1}\right)i_1+\left[R(1+\mu)\frac{d}{dt}+R_p\frac{d}{dt}+\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right]i_2-R_p\frac{d}{dt}i_3=0 \tag{7}$$

$$-\frac{\mu}{C_1}i_1+\left(R_p\frac{d}{dt}+\mu R\frac{d}{dt}+\frac{\mu}{C_1}\right)i_2-\left(L_p\frac{d^2}{dt^2}+R_p\frac{d}{dt}\right)i_3=0 \tag{8}$$

From the above simultaneous differential equations we obtain by differentiation and elimination the equation which each of the currents must satisfy:

$$\begin{aligned}
&\{L_1L_p[(1+\mu)R+R_p]\}D^4i\\
&+\left\{L_1RR_p+L_pR_1[(1+\mu)R+R_p]+L_1L_p\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)\right\}D^3i\\
&+\left\{L_1R_p\left(\frac{1}{C_{pg}}+\frac{1}{C_1}\right)+L_pR_1\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)\right.\\
&\left.+L_p\frac{1}{C_1}[(1+\mu)R+R_p]+R_1RR_p\right\}D^2i\\
&+\left\{R_1R_p\left(\frac{1}{C_{pg}}+\frac{1}{C_1}\right)+R_pR\frac{1}{C_1}+L_p\frac{1}{C_1}\frac{1}{C_{pg}}\right\}Di\\
&+R_p\frac{1}{C_1}\frac{1}{C_{pg}}=0.
\end{aligned} \tag{9}$$

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Suppose, for convenience, that this equation is written as:

$$(D^4+aD^3+bD^2+cD+d)i=0 \tag{10}$$

The general solution will be of the form:

$$i=Ae^{\alpha_1 t}\sin(\omega_1 t+\theta_1)+Be^{\alpha_2 t}\sin(\omega_2 t+\theta_2)$$

where A, B, θ_1, and θ_2 are arbitrary constants and α_1, α_2, ω_1, ω_2 are such that $\alpha_1+j\omega_1$; $\alpha_1-j\omega_1$; $\alpha_2+j\omega_2$; $\alpha_2-j\omega_2$ are the roots of (10) regarded as an equation in powers of D.

The following relations will hold between the roots and the coefficients of (10):

$$\left.\begin{aligned} 2(\alpha_1+\alpha_2)&=-a \\ 4\alpha_1\alpha_2+\Omega_1^2+\Omega_2^2&=b \\ 2\alpha_2\Omega_1^2+2\alpha_1\Omega_2^2&=-c \\ \Omega_1^2\Omega_2^2&=d \end{aligned}\right\} \tag{11}$$

where

$$\Omega_1^2=\alpha_1^2+\omega_1^2$$

and

$$\Omega_2^2=\alpha_2^2+\omega_2^2$$

It may be easily shown that if either α_1 or α_2 is zero, the following relation must hold among the coefficients:

$$c^2+a^2d-abc=0 \tag{12}$$

Conversely, if (12) is true then either α_1 or α_2 must be zero, for if we substitute for a, b, c, and d in (12), the values given by (11), and reduce, we obtain:

$$c^2+a^2d-abc=-\alpha_1\alpha_2\left[4(\alpha_2\Omega_1^2+\alpha_1\Omega_2^2)(\alpha_1+\alpha_2)+(\Omega_1^2-\Omega_2^2)^2\right] \tag{13}$$

Now it is seen by reference to (9) and (10) that a and c are always positive. Hence, $4(\alpha_2\Omega_1^2+\alpha_1\Omega_2^2)(\alpha_1+\alpha_2)$, which is equal to $a\,c$, must be positive and the coefficient of $\alpha_1\alpha_2$ in (13) is under no circumstances zero. Hence if c^2+a^2d-abc vanishes, either α_1 or α_2 must be zero. Suppose α_1 is zero.

From the first of the equations (11) it is seen that $\alpha_1+\alpha_2$ must be negative. Under these circumstances, therefore, oscillations of frequency ω_1 when once started will be maintained with unchanged amplitude. Oscillations of frequency ω_2, which occur

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when the circuit is given an electrical jar, will die out because α_2 is negative.

Therefore (12) regarded as an equation in R gives the value of R for which oscillations will just be maintained. This is the critical value, R_c, of the stabilizing resistance.

If in (12) we substitute for a, b, c, and d their values in terms of the circuit constants and clear of fractions, we obtain by placing $R=R_c$:

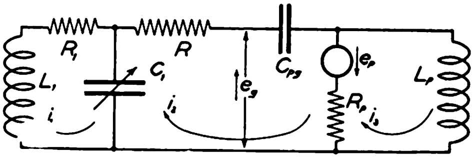

Fig. 2

$$L_1L_p[(1+\mu)R_c+R_p]\left\{R_1R_p\left(\frac{1}{C_{pg}}+\frac{1}{C_1}\right)+R_pR_c\frac{1}{C_1}+L_p\frac{1}{C_1}\,\frac{1}{C_{pg}}\right\}^2$$
$$+\left\{L_1R_cR_p+L_pR_1[(1+\mu)R_c+R_p]\right.$$
$$\left.+L_1L_p\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)\right\}^2R_p\frac{1}{C_1}\,\frac{1}{C_{pg}}$$

$$-\left\{L_1R_cR_p+L_pR_1[(1+\mu)R_c+R_p]+L_1L_p\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)\right\}$$
$$\left\{L_1R_p\left(\frac{1}{C_{pg}}+\frac{1}{C_1}\right)+L_pR_1\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)\right.$$
$$\left.+L_p\frac{1}{C_1}[(1+\mu)R_c+R_p]+R_1R_cR_p\right\}$$
$$\left\{R_1R_p\left(\frac{1}{C_{pg}}+\frac{1}{C_1}\right)+R_pR_c\frac{1}{C_1}+L_p\frac{1}{C_1}\,\frac{1}{C_{pg}}\right\}=0 \qquad (14)$$

In case R_1 is negligible, (14) may be reduced to:

$$R_p\left\{L_1R_p^2+L_p^2\frac{(1+\mu)^2}{C_1}\right\}R_c^2$$
$$+L_p\left\{L_1R_p^2\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)-L_1R_p^2\frac{\mu}{C_{pg}}+L_p^2\frac{(1+\mu)^2}{C_1C_{pg}}\right.$$

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$$+L_pR_p^2\frac{1+\mu}{C_1}\Big\}R_c$$

$$+R_pL_p^2\frac{1}{C_{pg}}\left\{L_p\frac{1+\mu}{C_1}-L_1\mu\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)\right\}=0 \qquad (15)$$

The Critical Frequency

The critical angular velocity, ω_c, is the value of ω_1, say, when $\alpha_1=0$.

When $\alpha_1=0$, we see from equations (11) that;

$$2\alpha_2=-a$$

$$2\alpha_2\omega_c^2=-c$$

Therefore,

$$\omega_c^2=\frac{c}{a}$$

Substituting for c and a, and remembering that R is adjusted to the critical value R_c:

$$\omega_c^2=\frac{1}{L_1C_1}\left[\frac{R_pR_c+L_p\dfrac{1}{C_{pg}}}{R_pR_c+L_p\left(\dfrac{1}{C_{pg}}+\dfrac{1+\mu}{C_1}\right)}\right] \qquad (16)$$

General Equations for R_c When the Plate Load is of Any Type

Considering merely the input characteristics of the vacuum tube without detailed reference to the plate circuit, we may consider the grid circuit of the tube to be of the form shown in

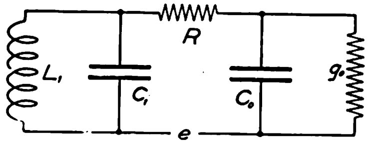

Fig. 3

Fig. 3. Here g_o is the input conductance of the tube and C_o is the input capacity. The circuit LC is assumed to be of negligible resistance.

The complex impedance which a current flowing under the action of an emf, e, would encounter is:

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$$z=R+\frac{g_o}{\omega^2C_o^2+g_o^2}-\frac{\omega C_oj}{\omega^2C_o^2+g_o^2}+\frac{\omega L_1j}{1-\omega^2L_1C_1}\,.$$

If any oscillations started in the circuit are just maintained after the removal of e, both the real and imaginary parts of z must be zero. Hence we must have:

$$R=R_c=\frac{-g_o}{\omega_c^2C_o^2+g_o^2} \qquad (17)$$

and

$$\frac{\omega_cL_1j}{1-\omega_c^2L_1C_1}=\frac{\omega_cC_oj}{\omega_c^2C_o^2+g_o^2}$$

or

$$\omega_c^2=\frac{C_o-L_1g_o^2}{L_1C_o(C_o+C_1)} \qquad (18)$$

Verification of Results

The circuit shown in Fig. 4 was used in checking (15). Oscillation was detected by heterodyning with the autodyne circuit A, which was very loosely coupled to the plate inductance, L_p. It was found that the sudden change in plate current caused by the commencement or cessation of oscillations was a more sensitive criterion of oscillation than was given by the autodyne circuit.

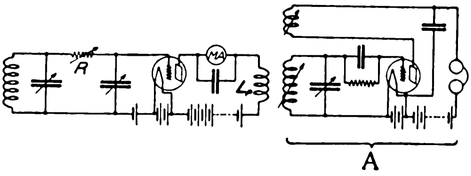

Fig. 4

The grid to filament capacity was made variable by means of a small condenser connected between the grid and filament. The total grid to filament capacity, C_{gf}, then, was the sum of the capacities of this condenser, the wiring, the socket, and the internal capacity of the tube.

The data for the curve shown in Fig. 5 was obtained by setting C_{gf} at some known value and then increasing R until oscillations ceased. These values of R_c were then plotted against the corresponding values of C_{gf} and the curve so obtained was extrapolated to give R_c when $C_{gf}=0$. The values of the critical resistance thus obtained were found to agree with the values calculated from (15).

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If we solve for R_c and ω_c in (15) and (16) and place these values in (17) and (18), we can solve these latter equations simultaneously for C_o and g_o. C_o and g_o as calculated by this method agreed exactly with the values calculated by means of Ballantine's formulas.[2]

Discussion

Fig. 5 shows the great change of R_c for a small change in grid to filament capacity when this capacity is small. The results given here may be made to serve as a basis for approximation formulas which will take account of the grid to filament capacity. Equations (17) and (18) may be made the basis of such formulas.

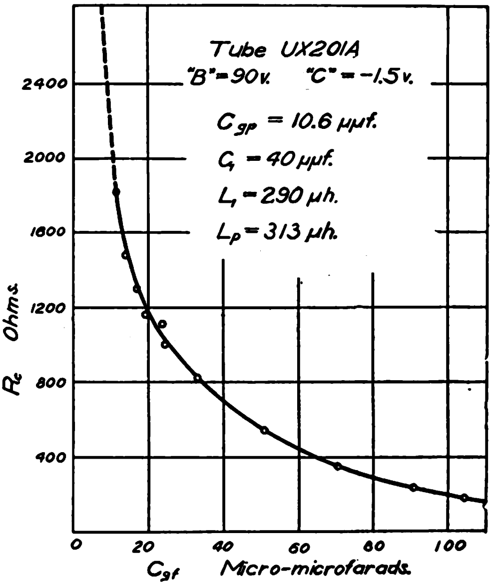

Fig. 5

If in equation (15), $R_c=0$, the critical condition is expressed by:

$$L_p\frac{1+\mu}{C_1}-L_1\mu\left(\frac{1}{C_{pg}}+\frac{1+\mu}{C_1}\right)=0. \tag{19}$$

In case C_1 is zero we have an Hartley oscillator circuit with no mutual inductance between the grid and plate circuits. The above equation then becomes:

[2] *Phys. Rev.*, 15, 409, 1920.

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$$L_p - \mu L_1 = 0. \tag{20}$$

This critical relation is in agreement with the observations of several writers.

The critical resistance could always be located to within a fraction of one per cent of its value. This fact and the great steepness of the curve in Fig. 5 suggest a method of measuring very small capacities. The unknown capacity would be placed between the grid and the filament of the vacuum tube. The change of the critical resistance thus produced by the added capacity would be a measure of this capacity. Such an application would, however, require consideration of the stability of operation of the tube.

Fig. 5 demonstrates the possibility of controlling regeneration by means of a condenser between the grid and the filament. If R remains fixed, we can, by varying C_{gf}, change the critical resistance R_c and cause it to approach R. When the circuit is not oscillating R must be greater than R_c. As R_c approaches R the regeneration will increase.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

THE PIEZO-ELECTRIC CRYSTAL OSCILLATOR*

By
J. WARREN WRIGHT
(Bureau of Engineering, Navy Department, Washington, D. C.)†

Summary—*The zero-angle quartz crystal, having electrodes in direct contact with its surface, is used to control the output of a typical oscillating crystal circuit.*

The crystal is first considered as a simple mechanical oscillator, and the required plate circuit adjustment for sustained vibrations of the crystal obtained by Miler's method on the basis of an assumed electrically equivalent crystal circuit.

The electrical equivalent of the crystal is then considered as the grid circuit of an oscillating vacuum-tube circuit and the equations for the frequency and condition for oscillation derived.

The effects of the tube and circuit upon the frequency of a crystal-controlled oscillator are then shown.

THERE are several very good articles[1] on the piezo-electric crystal resonator in the current literature, but little information has been made available, for general use, concerning the operation and theory of the piezo-electric crystal oscillator.[1,2] In this article an attempt is made to discuss the piezo-electric crystal oscillator from a theoretical and mathematical viewpoint.

We wish to examine just what takes place when a crystal is placed in a typical oscillating crystal circuit such as is shown in Fig. 1, and the circuit is adjusted so that sustained crystal-controlled oscillations result.

Let us first consider the crystal merely as an elastic substance placed between the plates *a* and *b*, Fig. 2. Suppose that pressure is applied to plate *a* (plate *b* assumed fixed), causing a slight deformation of the crystal, and then suddenly removed. Owing to its elastic properties, the crystal will tend to spring back to its former shape, but owing to its inertia the process will be carried too far, and thus a system of damped vibrations depending on the dimensions and physical properties of the quartz will be set up.

* Dewey decimal classification: R214. Original manuscript received by the Institute, April 30, 1928; revised manuscript received October 3, 1928. Published by permission of the Navy Department.
† Formerly of Naval Research Laboratory.
[1] For example, D. W. Dye, *Proc.* Phys. Soc. of London, **38**, part 5, 399, 1926; and W. G. Cady, PROC. I.R.E., **10**, 83; April, 1922.
[2] G. W. Pierce, *Proc.* Amer. Acad. of Arts and Sciences, **59**, 82, 1923.

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We will confine our attention to vibrations along the X axis[3] of the crystal, Fig. 2, and assume that the crystal acts like a simple mechanical system when set in vibration. Then if we let x be the displacement of the face of the crystal in contact with plate a along the X crystal axis, above or below the position assumed by this face when the crystal is not oscillating, we can write

$$(\alpha D^2+\beta D+\gamma)x=F=F_0 \cos \omega t \tag{1}$$

where $D=\dfrac{d}{dt}\cdot$ F is the applied force and it will be assumed that it affects only the damping term directly. Therefore such substitutions as are made in (2) and (5) hold only at or very close to the resonant frequency, where F and Dx are nearly in phase. α, β, and γ are the coefficients of inertia, damping, and restoration, respectively. They are constants of the quartz and its mounting.

It will be assumed, in all that follows, that we are dealing with vibrations along the X crystal axis and that the elastic limit of the quartz is never exceeded.

Equation (1) can take three forms, that is, in Equation (1), F can be (1) positive, (2) zero, or (3) negative. Cases (2) and(3) do not interest us here as they do not give sustained oscillations, and in solving Equation (1) we are only interested in the cases where oscillations are possible, and hence will omit the others.

Case (1a). F positive and just equal to the damping term. Equation (1) becomes

$$(\alpha D^2+\gamma)x=0 \tag{2}$$

The solution of (2) can be written

$$x=A \sin (\eta t+\rho) \tag{3}$$

where A is the amplitude of vibration, ρ is a constant, and η is the angular velocity.

An examination of Equation (3) shows that the period of vibration of the crystal, under these conditions, is given by

$$T=\frac{2\pi}{\eta} \tag{4}$$

[3] A. Crossley, Proc. I.R.E., **15**, 9; January, 1927.

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where T is the time in seconds for one complete mechanical oscillation.

Now $T=\frac{1}{f}$ where f is the frequency of vibration, hence $f=\frac{\eta}{2\pi}$.

Case (1b). F is positive and greater than the damping term. Under these conditions Equation (1) becomes

$$(\alpha D^2 - K\beta D + \gamma)x = 0 \tag{5}$$

where K is a factor depending on the magnitude of F and βDx.

Equation (5) states (under the assumption that F and Dx are in phase) that more energy is being supplied to the system than required to overcome its losses. The solution of (5), if we let $\delta_1 = \frac{K\beta}{\alpha}$ and $\eta^2 = \frac{\gamma}{\alpha}$ can be written as

$$x = A_1 e^{1/2\delta_1 t} \sin\left(\sqrt{\eta^3 - \tfrac{1}{4}\delta_1^2}\, t + \epsilon_1\right) \tag{6}$$

where A_1 is the original amplitude of vibration and ϵ_1 is a constant.

The period of vibration, in this case, can be shown to be

$$T_1 = \frac{2\pi}{\sqrt{\eta^2 - \frac{1}{4}\delta_1^2}} \tag{7}$$

and if this is expanded, by the use of Taylor's Theorem, we have

$$T_1 = \frac{2\pi}{\eta} + \frac{\pi\delta_1^2}{4\eta^3} + ---- - \tag{8}$$

An inspection of (8) shows that the period of vibration has increased owing to the fact that more energy is being supplied to the system than is needed to overcome the losses. The extra energy over and above that used to supply the losses, goes to build up larger oscillations. An inspection of (6) shows that, under these conditions, the oscillations would continue to increase in amplitude indefinitely. We can avoid this absurdity, in our case, by remembering that the energy must be supplied by the tube (see Fig. 1) and that, as the energy supplied is electrical, the characteristic curve of the tube employed serves to limit the increase in amplitude of the mechanical oscillations. Hence once the amplitude of vibration A_1 of (6) reaches this limit, there is

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no further increase in amplitude nor change in the period of the oscillations.

Experiments[4] have shown that when pressure is applied to a piece of piezo-electric quartz (zero-angle[3] cut considered here) a piezo-electric charge is developed on plates *a* and *b* which are in intimate contact with the quartz crystal. This is true whether the pressure is applied along the *X* or the *Y* axis. See Fig. 2.

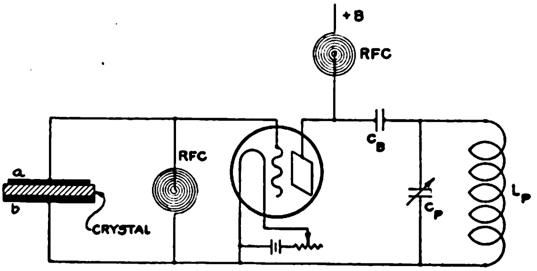

Fig. 1

It follows at once that if a crystal is caused to vibrate, these vibrations produce changes in stress in the crystal itself, and that a component of piezo-electric charge in phase with this stress will be developed on the electrodes *a* and *b*. The converse, that an oscillating charge will produce a mechanical vibration in a piezo-electric quartz crystal, is also true.

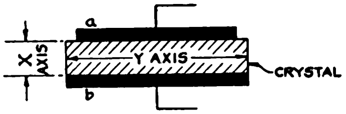

Fig. 2

Such a charge is due to the piezo-electric properties of the quartz crystal and it gives rise to an emf which can be expressed as

$$e = \frac{K'Q}{C} e^{\pm \tau t} \sin \omega t \qquad (9)$$

where $e^{\pm \tau t}$ is a damping term which becomes unity if τ is zero, Q is the piezo-electric charge, C is associated capacity, K' is a constant, and $\omega = 2\pi f$ where f is the frequency of vibration of the quartz crystal.

[4] This effect was first noted by the Curies. A recent detailed study of it was made by L. H. Dawson, *Phys. Rev.*, 29, 532, 1927.

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It is well to note by (9) that, for a given charge Q, the emf developed by the crystal is inversely proportional to the capacity associated with it. This capacity is the sum of the capacities of the crystal and holder, the capacity of the leads and the tube capacity. This capacity might be considered as a part of C', Fig. 3.

The frequency f is that of the mechanical vibration of the quartz, and is a function of the dimensions of the crystal and of the axis used. It is also a function of the input capacity of the vacuum tube used and the amplitude of mechanical vibration. In a properly ground zero-angle crystal there are three natural frequencies[3,5]; (a) that corresponding to the mechanical vibration of the crystal along the X axis, (b) that corresponding to the mechanical vibration along the Y axis, and (c) the so-called *coupling frequency.*

It would seem, then, that the frequency of a crystal oscillator might shift from one to another of the natural frequencies of the crystal. This is not the case, however, with a properly ground crystal, since the natural frequencies are separated far enough from each other so that tuning the plate circuit, Fig. 1, to produce oscillations at one frequency results in circuit conditions which are, in general, unfavorable to the production of the other two frequencies.

In order for the circuit of Fig. 1 to generate sustained crystal-controlled oscillations, assuming the crystal to impress an emf as given by (9) on the grid of the vacuum tube, the losses in the crystal must be supplied by the tube circuit. When this condition is satisfied the crystal continues to vibrate with constant amplitude, and owing to its piezo-electric properties it excites the grid of the tube with an undamped emf. This means that the damping term of (9) becomes unity for this condition.

Now in the case of the piezo-electric quartz crystal we have a mechanical vibration producing an electrical oscillation, and in general an electrical oscillation can be represented by an equation of the form

$$\left(LD^2+RD+\frac{1}{C}\right)q=e \qquad (10)$$

where $D=\dfrac{d}{dt}$, l is the emf, q is the charge, while L, R, and C are the inductance, resistance, and capacity, respectively.

[5] A. Hund, Proc. I.R.E., 14, 447; August, 1926.

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The similarity between (1) and (10) is marked and it has been shown[6] that an equivalent electrical system can be substituted for the mechanical one for the purpose of studying the frequency and conditions for oscillations in a circuit such as Fig. 1. However, it must not be assumed that in actual practice a crystal can be replaced by an equivalent electrical circuit and still afford the stability of frequency which is a feature of the crystal-controlled oscillator.

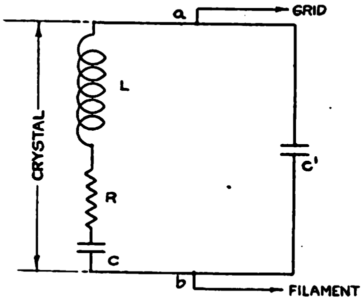

Fig. 3

Furthermore, it has been shown that the series chain of such an equivalent electrical system (Fig. 3) would have constants[6,7] given by

$$L = a\alpha \qquad R = b\beta \qquad C = c\gamma \qquad (11)$$

where α, β, and γ are as defined in (1), while L, R, and C are the inductance, resistance, and capacity, respectively, and a, b, and c are constants. C' is the capacity of the quartz condenser.[7]

Such an equivalent electrical system[1] is shown in Fig. 3 where it is assumed that the electrodes rest directly on the crystal.

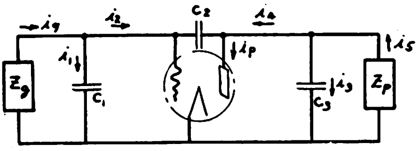

Fig. 4

Since we can replace a given crystal operating along a given axis by a fixed equivalent electrical circuit, our problem is to show the necessary conditions of adjustment of the plate circuit,

[6] Butterworth, *Proc.* Phys. Soc. of London, 271, 410, 1915.
[7] Van Dyke, K. S., Proc. I.R.E., 16, 742; June, 1928.

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Fig. 1, so that the resistance term of our equivalent electrical circuit is neutralized. This will also be the condition for the damping term of (1) to vanish, and hence if we can find such an adjustment sustained crystal-controlled oscillations are possible.

Miller[8] has shown that a vacuum tube will supply energy to the input circuit provided that the load in the plate circuit is inductive. Hence the condition for the damping term of (1) to vanish can be satisfied by tuning the plate circuit (Fig. 1), so that the resonant frequency of L_pC_p is greater than the crystal frequency.

Let us now attack the problem using Miller's[8] method, and determine the required plate circuit tuning so that the tube circuit will neutralize the resistance of the external input circuit.

We shall refer to Fig. 4 and assume that the vacuum tube has a linear characteristic, and that the grid is held at such a potential with respect to the filament that it does not draw current. In this figure Z_g represents the input impedance, Z_p represents the load in the plate circuit, and C_1, C_2, and C_3 are the inter-electrode capacities.

From an inspection of Fig. 4 and using Kirchoff's laws we can write the following equations

$$i_g = i_1 + i_2 \tag{12}$$

$$i = i_2 + i_4 \tag{13}$$

$$i_4 = i_5 - i_3 \tag{14}$$

$$i_1 = e_g j\omega C_1 \tag{15}$$

$$i^5 Z_p + \frac{i_3}{j\omega C_3} = 0 \tag{16}$$

$$\frac{i_3}{j\omega C_3} + \frac{i_2}{j\omega C_2} - \frac{i_1}{j\omega C_1} = 0 \tag{17}$$

and from vacuum-tube theory, assuming a linear characteristic,

$$\mu e_g - i_p r_p - i_5 Z_p = 0, \tag{18}$$

where μ is the amplification constant, e_g is the instantaneous a.c. grid voltage, and r_p is the internal resistance (plate to filament) of the vacuum tube.

[8] John Miller, Bureau of Standards Scientific Paper No. 351, 1919.

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Substituting (12), (13), (14), (15), (16), and (17) in (18) we have

$$e_g\left(\mu+j\omega C_1 r_p+\frac{r_p}{Z_p}+\frac{r_p}{Z_p}\,\frac{C_1}{C_2}+j\omega C_3 r_p+\frac{j\omega C_1 C_3 r_p}{C_2}+\frac{C_1}{C_2}+1\right)$$

$$=i_g\left(r_p+\frac{r_p}{j\omega C_2 Z_p}+\frac{r_p C_3}{C_2}+\frac{1}{j\omega C_2}\right) \tag{19}$$

Equation (19) can be expressed in terms of an impedance if we remember that $Z_g=\dfrac{e_g}{i_g}$, so we have

$$Z_g=\frac{r_p(C_2+C_3)+\dfrac{r_p}{j\omega Z_p}+\dfrac{1}{j\omega}}{\mu C_2+C_2+C_1+\dfrac{r_p}{Z_p}(C_1+C_2)+j\omega r_p(C_1C_2+C_1C_3+C_2C_3)} \tag{20}$$

and substituting $Z_p=R_p+jX_p$, equation (20) becomes

$$Z_g=\frac{R_p(C_2+C_3)+\dfrac{X_p}{\omega r_p}+j\left[X_p(C_2+C_3)-\dfrac{1}{\omega}-\dfrac{R_p}{\omega r_p}\right]}{\dfrac{R_p}{r_p}(\mu C_2+C_2+C_1)+(C_1+C_2)-\omega X_p(C_1C_2+C_2C_3+C_1C_3)+j\left[\dfrac{X_p}{r_p}(\mu C_2+C_2+C_1)+\omega R_p(C_1C_2+C_2C_3+C_1C_3)\right]}$$

This equation can be written in the form

$$Z_g=\frac{a+jb}{c+jd}=\frac{ac+bd}{c^2+d^2}+j\frac{bc-ad}{c^2+d^2}=R_g+jX_g \tag{22}$$

Here R_g is the resistance and X_g the reactance of the internal input impedance of the tube, and

$$a=R_p(C_2+C_3)+\frac{X_p}{\omega r_p}$$

$$b=X_p(C_2+C_3)-\frac{1}{\omega}-\frac{R_p}{\omega r_p}$$

$$c=\frac{R_p}{r_p}(\mu C_2+C_2+C_1)+(C_1+C_2)-\omega X_p(C_1C_2+C_2C_3+C_1C_3) \tag{23}$$

$$d=\frac{X_p}{r_p}(\mu C_2+C_2+C_1)+\omega R_p(C_1C_2+C_2C_3+C_1C_3)$$

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An examination of the resistance term of (22) shows that if X_p is positive, corresponding to an inductive load in the plate circuit, R_g becomes negative when

$$\frac{\mu X_p}{\omega r_p} > \frac{X_p^2}{r_p}[\mu(C_2+C_3)+C_2] + \frac{R_p^2}{r_p}[\mu(C_2+C_3)+C_2] + R_pC_2 \quad (24)$$

Hence if we make R_g negative by the use of the inequality (24) it means that the tube circuit is regenerating, and if we make R_g sufficiently negative so that the tube supplies the losses of the external input circuit, then oscillations are possible.

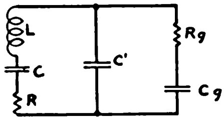

Fig. 5

This means that the tube circuit supplies the losses in the external input circuit and neutralizes the effect of the damping terms of (1) and (9). Hence, we have proved[9] that if the load in the plate circuit is made inductive, the resistance, and therefore (if we assume the use of the equivalent electrical circuit to be justified) the damping of the crystal can be overcome by the tube's supplying energy to the input, provided that the resistance is not too great.

An examination of the reactance term of (22) shows that X_g is a capacity reactance, and that the input impedance of the tube, under these conditions, can be represented by a resistance R_g in series with a capacity C_g. The values of this resistance and capacity are given by

$$R_g = \frac{ac+bd}{c^2+d^2} \quad (25)$$

and

$$C_g = \frac{c^2+d^2}{\omega(ad-bc)} \quad (26)$$

where a, b, c, and d are as given in (23).

The manner in which the electrical equivalent of the crystal is connected to the input of the vacuum tube is shown in Fig. 5. Here that portion of the circuit containing R_g and C_g in series

[9] This proof is in accord with Miller's, and is given here because it is felt that few engineers are familiar with his excellent paper.

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represents the input impedance of the tube while the remainder represents the electrical equivalent of the crystal.

If we refer to Fig. 4 and to (25), and assume that R_g is negative, we note that the power supplied by the tube to the input circuit is given by

$$i_g^2 R_g = P_g \tag{27}$$

Now refer to Fig. 5 and assume[8] that a current i flows in that portion of the circuit representing the series chain of the crystal. If the reactance of C_g is large compared to the value of R_g, this current will divide between C' and C_g in proportion to their capacities. That part flowing through the series circuit containing R_g and C_g is the current referred to in Fig. 4 as i_g and can be expressed as

$$i_g = i \frac{C_g}{C' + C_g} \tag{28}$$

The power dissipated in the crystal is given by

$$P_C = i^2 R \tag{29}$$

If we substitute (28) in (27), and add the resultant power equation to (29) we have

$$P = i^2 \left[R - \left(\frac{C_g}{C' + C_g} \right)^2 R_g \right] \tag{30}$$

where the negative sign indicates that the tube is supplying power to the input.

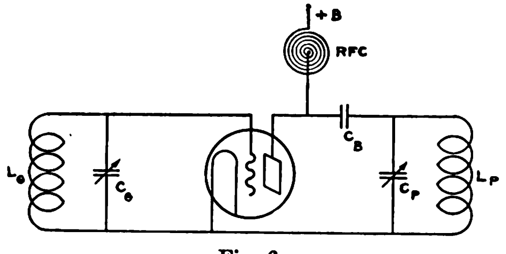

Fig. 6

When adjustments are made so that (30) becomes zero, sustained crystal-controlled oscillations are possible. It is well to note, however, that the frequency of these oscillations may differ if the tube is replaced by one having different interelectrode capacities, or if the plate circuit is retuned. These facts are

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made clear by a study of Fig. 5 and (26). The change in the frequency of the crystal-controlled oscillations is very small, but may be important in the case of crystal-controlled oscillators which are to be used as frequency standards.

We have shown, assuming the use of the equivalent electrical circuit to be justified, that the plate circuit can be tuned so that the tube circuit will supply electrical energy to the input circuit, and, owing to the piezo-electric properties of the quartz crystal, maintain the mechanical vibrations of the crystal. These mechanical vibrations produce piezo-electric charges, and the tube is therefore excited by an undamped emf of the same frequency as that of the mechanical vibration of the crystal.

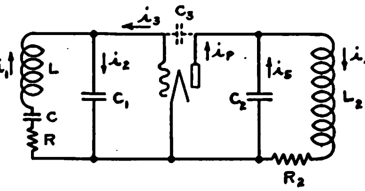

Fig. 7

We shall next consider the crystal-controlled oscillator as an oscillating vacuum-tube circuit, for it is well known that an inductance shunted by a capacity can be substituted for a crystal in an emergency, and that the circuit will then function with only slight changes in adjustments. This circuit is shown in Fig. 6 and is the so-called "Tuned-Grid Tuned-Plate Circuit."[10] It does not represent the electrical equivalent of the crystal[7] oscillator closely enough to be of interest here, so we shall replace the grid circuit of Fig. 6 by the electrical equivalent of the crystal as shown in Fig. 3. The resultant circuit is shown in Fig. 7.

Let us refer to Fig. 7 and assume, for simplicity, that the grid does not draw current, that the tube has a linear characteristic, and consider only small oscillations. In addition assume that the resistance is small compared to the reactance.

Now express the various currents of Fig. 7 in terms of the product of a voltage and an admittance.

Thus in the grid circuit

$$i_1 = e_g\left(g_1 - \frac{j}{X}\right) \tag{31}$$

[10] J. W. Wright, Proc. I.R.E., **16**, 1113; August, 1928.

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$$i_2 = e_g j\omega C_1 \tag{32}$$

$$i_3 = (e_p - e_g) j\omega C_3 \tag{33}$$

$$i_1 = i_2 + i_3 \tag{34}$$

where $X = \left(\omega L - \frac{1}{\omega C}\right)$ and $g_1 = \frac{R}{X^2}$ e_g and e_p are the instantaneous alternating grid and plate voltages, respectively.

Substituting (31), (32), and (33) in (34) and solving for e_p,

$$e_p = \frac{g_1 + j\left\{\omega(C_1 + C_3) - \frac{1}{X}\right\}}{j\omega C_3} \tag{35}$$

Similarly in the plate circuit

$$i_4 = e_p \left(g_2 - \frac{j}{\omega L_2}\right) \tag{36}$$

$$i_5 = e_p j\omega C_5 \tag{37}$$

$$i_p = -e_g G \tag{38}$$

$$i_p = i_3 + i_4 + i_5 \tag{39}$$

where $g_2 = \frac{R_2}{\omega^2 L_2^2}$ and G[11] is the mutual conductance of the plate circuit.[12]

Substituting (33), (36), (37), and (38) in (39) we obtain

$$e_g G + e_p(g_2 + jb_2) - e_g j\omega C_3 = 0 \tag{40}$$

where

$$b_2 = \left\{\omega(C_2 + C_3) - \frac{1}{\omega L_2}\right\}$$

Now substitute the value of e_p as given by (35) in (40), eliminate the voltage terms, and clear of fractions

$$j\omega C_3 G + g_1 g_2 + jb_1 g_2 + jb_2 g_1 - b_1 b_2 + \omega^2 C_3^2 = 0 \tag{41}$$

where

$$b_1 = \left\{\omega(C_1 + C_3) - \frac{1}{X}\right\}$$

[11] L. A. Hazeltine, Proc. I.R.E., 6, 63; April, 1918.
[12] Van der Bijl, "Thermionic Vacuum Tube," McGraw-Hill, 1918.

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Generally g_1 g_2 and $\omega^2 C_3^2$ are very small, and if we neglect them and equate the remaining real terms of (41) to zero in order to obtain the frequency of oscillation, we have

$$b_1 b_2 = 0 \tag{42}$$

If the values of b_1 and b_2 are substituted in (42) and this equation solved for ω^2 it is found that there are two possible frequencies given by

$$\omega_1{}^2 = \frac{1}{L\dfrac{C(C_1+C_3)}{C+C_1+C_3}} \tag{43}$$

and

$$\omega_2{}^2 = \frac{1}{L_2(C_2+C_3)} \tag{44}$$

Here $\omega_1{}^2$ gives the frequency of the oscillations which are determined by the constants of the grid circuit. This frequency can be seen, by an inspection of (43), to be lower than the natural frequency of the crystal and mounting which is given by

$$\omega^2 = \frac{1}{L\dfrac{CC'}{C+C'}} \tag{45}$$

where the symbols are the same as used in Fig. 3.

This is seen to be due to the fact that the capacity C_1 (Fig. 7), is approximately equal to the sum of the grid-filament capacity of the tube and the shunt capacity of the crystal, C', of Fig. 3. Therefore the capacity term of (43) is greater than the corresponding term of (45). The result is that the frequency as given by (43) is lower than that given by (45).

It is very interesting to note that the input capacity of the tube enters into the frequency equations. This is in accordance with the predictions made after examining Fig. 5, and it may account for certain variations in the frequency of crystal oscillators, which are used as standards of frequency, when the tubes are changed.

The condition for the starting of oscillations is obtained from the imaginary terms of (41) by writing

$$\omega C_3 G + b_1 g_2 + b_2 g_1 = 0 \tag{46}$$

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Substituting the values of b_1, b_2, g_1, and g_2 in (46), solving for G and neglecting certain very small quantities the condition for oscillations of the frequency given by (43) to occur is that

$$G=\frac{RC^2C_4(LC_4-L_2C_5)}{LL_2C_3(C-C_4)^2} \tag{47}$$

where $C_4=\dfrac{C(C_1+C_3)}{C+C_1+C_3}$ and $C_5=(C_2+C_3)$.

Now G cannot become negative, so $LC_4 \geqq L_2C_5$, but because $(C-C_4)^2$ is so very small there is practically only one adjustment which satisfies the condition that the value of G given by (47) must not be negative or greater than that obtainable from the vacuum tube. In addition the oscillations will be stronger when G is a minimum,[11] which occurs when

$$LC_4=L_2C_5 \tag{48}$$

Similarly the condition for the starting of oscillations of a frequency as given by (44) can be shown to be

$$G=\frac{R_2C_5\,(L_2C_5-LC_4)}{L_2C_3\,(C+C_1+C_3)} \tag{49}$$

where $C_4=\dfrac{C\,(C_1+C_3)}{C+C_1+C_3}$ and $C_5=(C_2+C_3)$.

In (49) the minimum value of G occurs when

$$L_2C_5=LC_4 \tag{50}$$

A comparison of (48) and (50) shows that the strongest oscillations occur when ω_1^2 is equal to ω_2^2. For all practical purposes the circuit of Fig. 7 can be considered to oscillate at only one frequency.

These conclusions agree with experiment, for it is well known that in tuning a crystal oscillator, Fig. 1, the load in the plate circuit should be adjusted so that its reactance is inductive, and then, when oscillations start, adjusted for maximum output, which is obtained when the plate circuit load is approximately in resonance with the crystal frequency.

We also know, from experimental evidence, that the emitted frequency of a crystal-controlled oscillator, Fig. 1, depends to a certain extent upon the circuit and tube as well as upon the

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dimensions of the crystal. If a given crystal and holder are used, in a circuit such as shown in Fig. 1, to control the output of a commercial type UX201A tube, the frequency of the crystal-controlled output will be higher than if the same crystal is used to control the output of a Navy type CW1818A tube connected to an exactly similar circuit. This is believed to be due to the fact that the input capacities of the two tubes are different. This change in the frequency of the output could have been predicted from an examination of (43).

Another effect, which has not been mentioned so far and which causes changes in the generated frequency of a crystal oscillator, is that of temperature and its relation to the frequency of vibration of a quartz crystal. An increase in the temperature of a crystal causes the constants α, β and γ to change in such a manner that the generated frequency of the crystal-controlled oscillator is lowered. A decrease in temperature causes an increase in the frequency of the oscillator.

The effect of temperature upon the output frequency also depends on which axis of the zero-angle crystal is being used. If the frequency corresponds to vibrations of the crystal along the X axis, the temperature coefficient of frequency for the crystal, in a Navy type holder, is approximately twenty-five parts in a million per degree C. The temperature coefficient of frequency for Y axis vibrations is approximately fifty parts in a million per degree C., assuming the crystal is mounted in a Navy type holder.

The above discussion applies particularly to the case where the output of the crystal-controlled oscillator is desired to be of the same frequency as the crystal. In certain cases, however, we desire the fundamental crystal frequency and as many harmonics of this frequency as possible (as in the crystal-controlled calibrator). This can be accomplished by tuning the plate circuit by means of an inductance having a large L/C ratio. The harmonics are generated in the electrical circuit, and it should not be assumed that the crystal itself vibrates at the harmonic frequencies. In case it did, the harmonics would not necessarily be integral multiples of the fundamental, whereas the observed harmonics of a crystal-controlled calibrator are integral multiples of the fundamental crystal frequency.

When a crystal oscillator is used to control the frequency of the usual type of transmitter, and the emitted frequency is

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an integral multiple of the crystal frequency, it does not follow that the desired harmonic present in the output of the crystal oscillator is picked up and amplified. On the contrary, except when the transmitter is operating at the crystal frequency, the crystal oscillator output excites a vacuum-tube frequency multiplier, which has its negative grid bias so adjusted and its plate circuit load so tuned that the desired frequency, which must be an integral multiple of the crystal frequency, is generated in the multiplier circuit.

Recent experiments have also shown that a crystal can be used to control the output frequency of an oscillator of the *relaxation* type so that crystal-controlled oscillations which are sub-harmonics of the fundamental crystal frequency can be obtained.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

FADING CURVES AND WEATHER CONDITIONS*

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Summary—*Sunset fading curves from Station KDKA were made at Morgantown with a Shaw Recorder during April and May, 1927. The curves taken on fine days show more irregularity during the daylight hours than those taken on cloudy days. It was found that the signal strength from KDKA during the dark hours sometimes fell far below the daylight strength. A falling curve indicates clearing weather for the next day while a rising curve is followed by cloudy weather or rain.*

During the winter months, KDKA, the Pittsburgh Station of the Westinghouse Electric and Manufacturing Company, is very seldom in continuous operation through the sunset period. It is therefore possible to get only four curves a month showing the variation of strength in radio reception

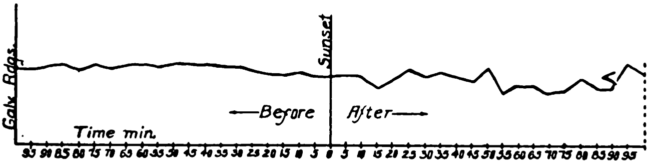

Fig. 1—KDKA at Morgantown; Generally Cloudy Days.

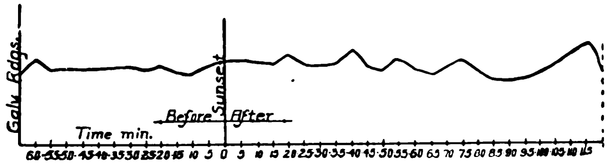

Fig. 2—KDKA at Morgantown; Generally Clear Days.

for an hour before and an hour after sunset. In the spring, however, on account of the longer days, it is comparatively easy to make records day after day throughout the sunset period. During April and May, 1927, thirty curves were taken on a Shaw Recorder[1] throughout the sunset period. Eight of these curves were *undisturbed*, that is, they were fairly even during the period before sunset. Twenty-two of the curves were irregular during the daylight hours. Four of the undisturbed curves were taken on clear days and four on cloudy days. Of the *disturbed* curves

* Dewey decimal classification: R113.1. Original manuscript received by the Institute, October 30, 1928.

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seventeen were taken on clear days and only five on cloudy days. This plainly shows the preponderance of irregularity on fine days.

The *undisturbed* curves were taken on April 4, 24, 1927, and May 2, 6, 7, 9, 11, 26, 1927. The *disturbed* curves were taken April 6, 11, 12, 13, 19, 20, 21, 22, 23, 25, 28, 30, 1927, and May 1, 3, 4, 5, 10, 12, 13, 14, 17, 27, 1927.

The galvanometer readings on the first eight curves were averaged at five-minute intervals measured from sunset as zero. The resulting curve is shown in Fig. 1. *Here it was noticed for the first time that the night reception from KDKA could fall below and remain below the daylight strength.* The twenty-two *disturbed*

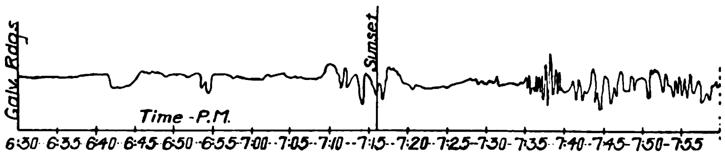

Fig. 3—Sunset Curve, KDKA, at Morgantown; Typical Cloudy and Cool. May 12, 1927.

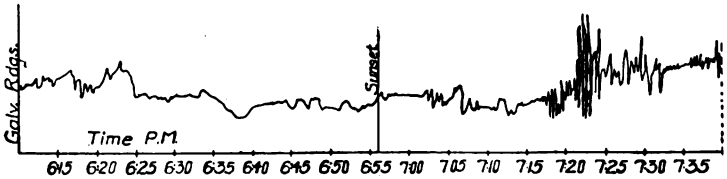

Fig. 4—Sunset Curve, KDKA, Morgantown; Typical Clear and Cool. April 23, 1927.

curves were averaged in the same way and the resulting curve is shown in Fig. 2. Fig. 3 shows a curve taken on the Shaw Recorder in which the night reception consistently falls below the day strength. Fig. 4 shows a curve in which the reception increases after sunset.

In the remaining figures, the upper curve in each case is that obtained on the Shaw Recorder while the lower curve shows the signal intensity obtained by averaging over five-minute intervals. That is, the whole curve is divided into five-minute intervals measuring from sunset or zero. The area of each interval is taken with a planimeter and the average ordinate calculated. The square root of the average ordinate gives the intensity

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on an arbitrary scale. Fig. 5 shows an abrupt increase of intensity after sunset; while Fig. 6 shows a slight increase about an

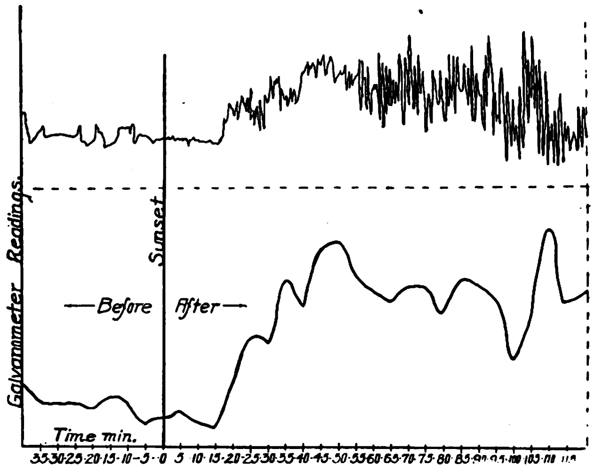

Fig. 5—Sunset Curve, KDKA, at Morgantown; Clear and Cool. April 11, 1927.

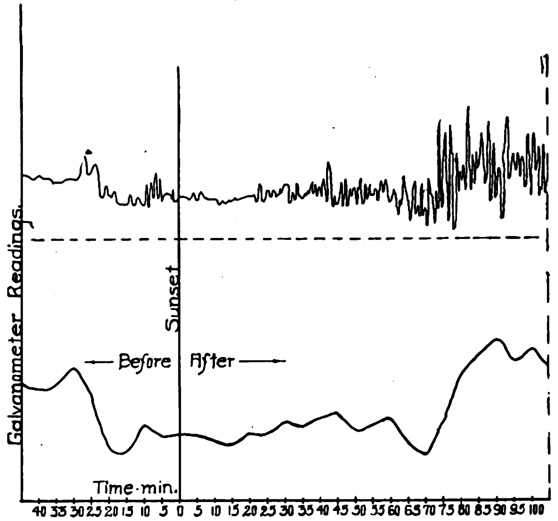

Fig. 6—Sunset Curve, KDKA, at Morgantown; Misting Rain. May 11, 1927.

hour after sunset. All the other graphs shown (Figs. 7 to 11) indicate a great reduction in signal strength after sunset.

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Attempts were made to continue these observations during the summer and winter of 1927–28, but heavy static in the sum-

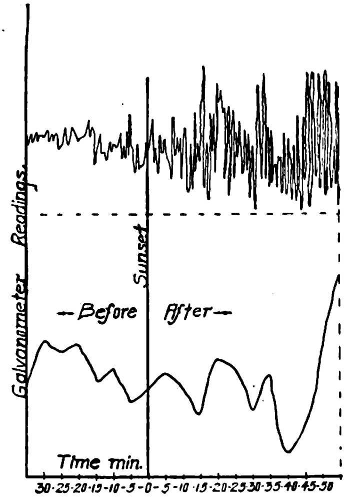

Fig. 7—Sunset Curve, KDKA, at Morgantown; Cool, Sunshine. April 6, 1927.

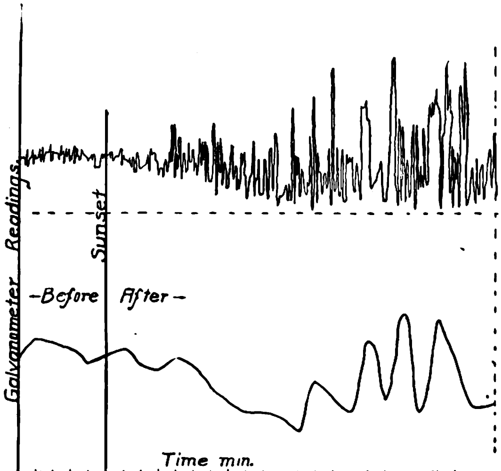

Fig. 8—Sunset Curve, KDKA, at Morgantown; Very Hazy, Temperature Pleasant. May 5, 1927.

mer of 1927 and interference from other stations during the winter prevented any continuous observations. However, the

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summer of 1928 was singularly free from static and observations made then show that weather conditions have a decided effect upon the signals from KDKA. So much so that it is even possible

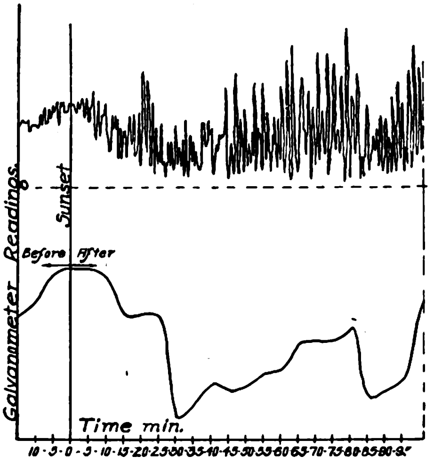

Fig. 9—Sunset Curve, KDKA, at Morgantown; Clear and Cool. May 13, 1927.

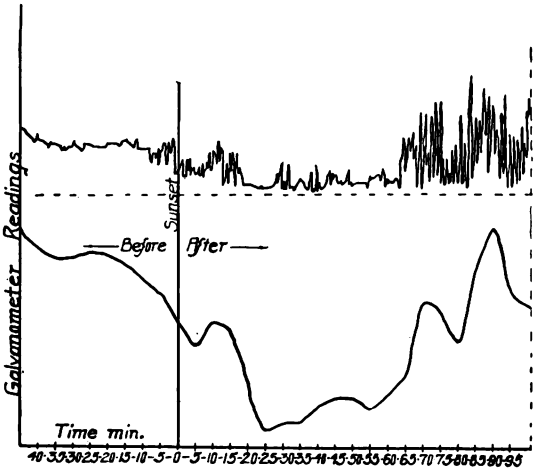

Fig. 10—Sunset Curve, KDKA, at Morgantown; Light Broken Clouds, Cool. May 14, 1927.

to foretell the weather one day ahead by the form of the fading curves. If the curve continues to rise after sunset, a wet or cloudy day is indicated; if the curve falls, the weather will tend

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to clear. Thus Fig. 6 (upper curve) taken on a cloudy day shows that the following day, May 12, 1928, would also be cloudy at Morgantown. The curve of Fig. 5 taken on April 11,

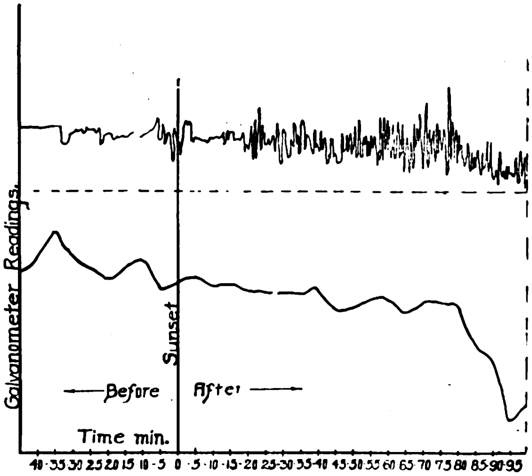

Fig. 11—Sunset Curve, KDKA, at Morgantown; Typical Cloudy and Cool. May 12, 1927.

1927 indicated that April 12 would be cloudy; April 12 was a cloudy day.

From Figs. 4 to 11, the captions "Galvanometer Readings" at the left refer to the upper curve only. The lower curve is the intensity of the signal averaged over five-minute intervals.

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DETECTION CHARACTERISTICS OF THREE-ELEMENT VACUUM TUBES*

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Summary—*The change of grid potential in a grid-leak grid-condenser detector can be determined by considering a fictitious "rectified voltage" acting in series with the grid resistance. This equivalent voltage is inversely proportional to the tube "voltage constant" v, which has the value* $v = 2\, R_g/(dR_g/dE_g)$*, and can be readily measured by an a.c. resistance bridge.*

The rectifying action of different tubes can be compared on the basis of the respective voltage constants at grid resistances inversely proportional to the size of grid condenser. Tubes are then compared under conditions of equal detector distortion, and the change of grid potential is inversely proportional to the voltage constants.

The voltage constant of ordinary vacuum tubes at first drops rapidly as the grid resistance is increased, but soon flattens out and becomes constant at grid resistances above 50,000 to 150,000 ohms.

The highest audio frequency that can be satisfactorily reproduced with the detector adjusted to full sensitivity is inversely proportional to the grid resistance at the lower end of the flat part of the $v - R_g$ *characteristic.*

It was found that tubes of the same type had uniform detection characteristics, that age, use, plate voltages between 16 and 122, and filament voltage (above the minimum necessary to give electron saturation) had little or no effect on the rectifying ability of high vacuum tubes at a given grid resistance in the useful range of operation.

Detector tubes best suited to resistance coupling are types 102-D, 240, and 200-A, because of their high μ *and low v. The best detector from the point of view of power output and quality is the 227 type because of its low voltage constant and high amplification. Next in merit comes a group composed of types 112-A, 226, and 12. The sensitivity of the 200-A alkali vapor tube is substantially that of the corresponding high vacuum tube with the same* μ*.*

DETECTOR CONSTANTS AND THEIR MEASUREMENT

IN the grid-leak grid-condenser detector the radio-frequency signal voltage is rectified in the grid circuit of the detector tube. This rectified grid current in flowing through the grid leak-condenser impedance produces a voltage drop which affects the plate circuit of the tube by amplifier action. In a previous paper[1] it was shown that the rectified current could be considered as produced by a series of fictitious generators acting between the grid and filament in series with the dynamic grid resistance. The voltage of these generators, conveniently called the "rectified

* Dewey decimal classification: R134. Original manuscript received by the Institute, September 5, 1928.

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voltage," depends only upon the signal voltage and a single tube constant.

Detector problems can accordingly be conveniently analyzed with the aid of the equivalent detector circuit of Fig. 1, in which the rectified grid voltage E_r acts in series with the grid resistance R_g, and the grid leak-condenser combination shunted by the tube input capacity C_g' to the rectified current. The voltage produced across the grid leak-condenser impedance by the rectified grid voltage is the change of grid potential which affects the plate current by amplifier action.

Two detector constants suffice to give the complete performance of the detector. The first of these, the grid resistance R_g, is the reciprocal of the slope of the grid current-grid voltage characteristic at the operating point, and is required in setting up the

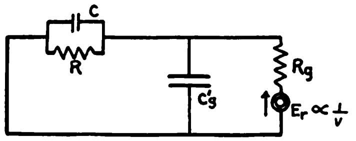

Fig. 1—Equivalent Circuit of Grid-Leak Grid-Condenser Detector.

equivalent detector circuit. The second constant determines the rectified grid voltage, and depends upon the grid resistance and the rate of change of grid resistance with grid voltage at the operating point. This constant has the dimension of a voltage, and so is called the voltage constant. It is defined by the relation

$$\text{Detector voltage constant} = v = \frac{2R_g}{\dfrac{dR_g}{dE_g}}$$

The rectified grid voltage is inversely proportional to v, so that a small voltage constant is desired. In the important case of a signal carrier wave of amplitude E_s modulated to a degree m, the modulation frequency component of rectified voltage has the amplitude mE_s^2/v.

To measure v one determines R_g first at the operating grid potential, and then at grid voltages above and below this operating point by a small amount ΔE_g, thus giving dR_g/dE_g. These three measurements of grid resistance give the entire detector performance, and can be easily and rapidly carried out with the alternating-current bridge arrangement described in the previous paper.

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In using the bridge method of measuring v and R_g it is easier to manipulate the apparatus when the 1000-cycle input to the bridge is large, and when the grid-voltage increment ΔE_g is considerable, but increasing these quantities beyond a certain point introduces errors. A study was accordingly made to determine the most satisfactory values to use in making measurements.

It was found that the observed values of v become progressively lower than the true values as ΔE_g is increased, and for a given ΔE_g the error is greater for tubes with a small voltage constant. Mathematical analysis based on observed detector characteristics shows that over the range of grid voltages for which v is substantially independent of grid resistance (see Fig. 2 for example) values of ΔE_g such that $\Delta E_g/v=0.2$ lead to an error of less than 3 per cent. Larger voltage increments than this rapidly increase the error. Over the part of the tube characteristic where v decreases with increase of grid resistance the allowable $\Delta E_g/v$ is about half that permissible over the flat portion. As most tubes have a voltage constant v lying between 0.2 and 0.5 volt over the flat portion, and more over the dropping part, values of 0.05 volt and less for ΔE_g should give satisfactory results, and experiments show this to be the case.

The effect which the bridge input voltage has on the dynamic grid resistance as measured by the bridge depends primarily upon the purity of wave form of this voltage. The indications are that with a pure sine wave, the bridge input could be at least 0.1 volt effective in most cases. With harmonics present the bridge input must be smaller in proportion. The measurements reported in this paper were made using power from a master-oscillator power-amplifier vacuum-tube generator, and with the outfit loaded fairly heavily a bridge input voltage of 0.025 effective was about the maximum allowable because of harmonics. Errors due to excessive bridge inputs can be readily detected by noting whether or not the balance is affected by lowering the input.

Basis for Comparing Detectors

In comparing different grid-leak grid-condenser detectors it is necessary to separate the rectifying and the amplifying action. The change of grid potential is given by the equivalent circuit of Fig. 1 and is determined by the rectifying action in the grid, but

[1] F. E. Terman, "Some Principles of Grid-Leak Grid-Condenser Detection," Proc. I. R. E., **16**, 1384; October, 1928.

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the effect which this has on the plate current is purely an amplifier problem.

The criterion for comparing the rectifying action of different detectors when used with the same size grid condenser is the value of the voltage constant v at a given grid resistance. *The change of grid potential produced by a given signal is inversely proportional to the voltage constant v when the grid resistance R_g and the size of grid condenser are fixed.*[2]

It is well-known that the higher modulation frequencies do not give as great a change of grid potential as the lower frequencies, and the magnitude of this effect depends upon the grid resistance at the operating point and the size of grid condenser.[3] *Where the same grid condenser is used, detectors compared at the same value of grid resistance will be compared under conditions of equal distortion.*

The best size of grid condenser is approximately ten times the effective grid-filament tube capacity to radio frequencies, so will tend to vary with different types of tubes. Detectors with different grid-condenser capacities must be compared under conditions which give the same distortion at the higher modulation frequencies, *and so are compared on the basis of the voltage constant v at grid resistances inversely proportional to the grid condenser capacity.* This gives a comparison under conditions of equal distortion, and under such circumstances the change of grid potential produced by the signal is inversely proportional to the voltage constant, so that *the tube with the smallest v gives the greatest change of grid potential.*

The overall performance of a detector depends upon the change of grid potential, the criterion for which has already been given, and the amplification of this grid voltage. Unfortunately, there is no simple physical quantity that correctly gives a figure of merit for the entire detector action, and for the present the two functions will be considered separately.

[2] This statement and the similar italicized phrases in the next two paragraphs assume that the grid-leak resistance is so high as to have negligible effect on the voltage drop across the grid leak-condenser impedance. The exact statement is that *for equal* distortion the conductance of the grid resistance and the grid-leak resistance in parallel should be proportional to the effective grid-condenser capacity. The conductance of the grid-leak resistance is ordinarily so small that for most uses the exact rule is not necessary.

[3] For an exact analysis see Terman, l.c.; or Ballantine, "Detection by Grid Rectification with the High-Vacuum Triode," Proc. I. R. E. **16**, 593; May, 1928.

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Results of Tests on High-Vacuum Detectors

General. Tests of numerous types of tubes under a wide variety of conditions *in every instance* gave a relationship between voltage constant v and grid resistance R_g of the general form shown in Fig. 2, in which with increasing grid resistance the voltage constant rapidly drops to a substantially constant value that continues up to the limit of the measuring equipment used, which was about two megohms. The principal differences observed between individual tubes, types of tubes, and battery potentials were in the value of voltage constant over the flat part of the curve, the grid resistance where the curve flattened out, and the grid voltage required to give a specified grid resistance.

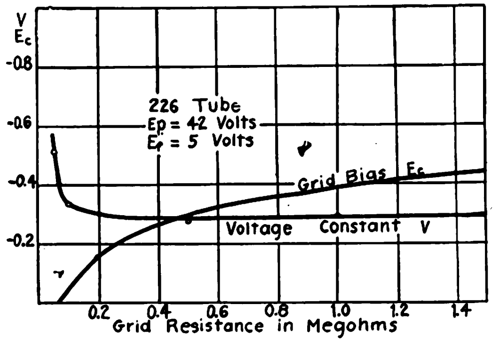

Fig. 2—Rectifying Characteristics of Typical Detector Tube.

In considering the significance of Fig. 2, it is to be remembered that for good quality telephone reception with standard tubes the grid resistance must be between 100,000 and 200,000 ohms. The exact value to use depends upon the allowable distortion at the high modulation frequencies and can be computed with the aid of the equivalent detector circuit of Fig. 1. In code reception, grid resistances several times the values most suitable for telephone reception are preferable. The important operating range of the detector is hence at grid resistances in the range from 100,000 to 500,000 ohms. The essential points to observe in comparing the detector action of different tubes are (1) the value of voltage constant over the flat part of the characteristic, and (2) the grid resistance at which the characteristic becomes flat.

Comparison of the Detector Action of Tubes of the Same Type under the Same Conditions. It was found that the $v-R_g$ characteristic of different tubes of the same type, operated at the same

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plate and filament voltages, was consistently uniform. The value of voltage constant v over the flat part of the characteristic for an individual tube practically never departs over 15 per cent from the average value of all tubes.

The principal individual difference between tubes is in the value of grid voltage required to give a given grid resistance. There were also some differences between tubes in the grid resistance at which the characteristic became flat, and in the characteristic before the flat portion was reached.

A typical comparison of several tubes of the same type is given in Fig. 3. The detection characteristics of individual tubes appear to be about as uniform as plate resistance, plate current, etc. This was found true for all types of tubes tested.

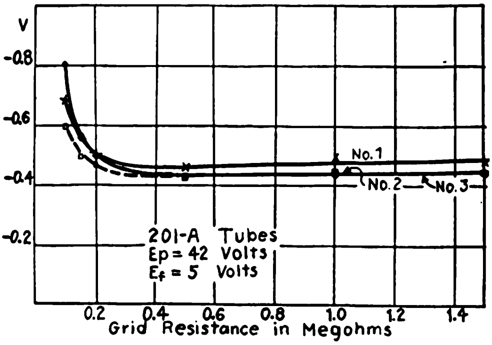

Fig. 3—Comparison of Three 201-A Tubes.

Effect of Filament Voltage. Substantially the only effect of changing the filament voltage of a tube is to alter the grid voltage at which a given grid resistance is obtained, this grid voltage being more negative at the higher filament temperatures. *The voltage constant v at a given grid resistance is roughly independent of filament voltage until the electron emission has dropped to the vicinity of the saturation value.* The results presented in Fig. 4 are typical of all tubes. In general the part of the characteristic before the flat part is much more influenced by electron emission than is the flat portion, and as the filament voltage is reduced the effect is noticed at the lower values of R_g before becoming evident at high grid resistances. The minima of v seen in Fig. 4 are found in all tubes.

Effect of Plate Voltage. Variations in plate voltage between about 16 and 122 volts have practically no effect on the value of voltage con-

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stant v at a given grid resistance provided this grid resistance is on the flat part of the $v-R_g$ *characteristic.* If on the dropping part, the voltage constant decreases slowly with increase of plate voltage. The results of Fig. 5 are typical. For the tube shown, the curve for $R_g=100{,}000$ ohms is the only one not on the flat part. The principal effect of increasing the plate voltage is to make a more positive grid necessary for a given grid resistance.

At very low plate voltages the detector voltage constant is somewhat erratic in its behavior, but is in general rather small, indicating good rectification. Some tubes, such as the 227 heater tube will actually rectify about as well with zero plate volts as at any voltage.

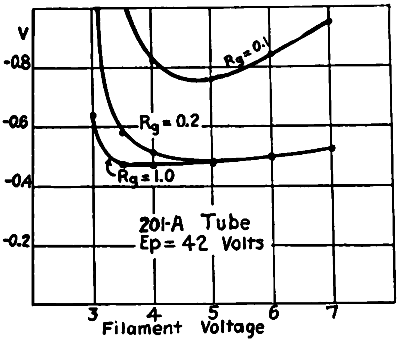

Fig. 4—Effect of Filament Voltage on Detector Voltage Constant.

Effect of Age. Age or use seems to have no effect on the $v-R_g$ characteristic of detectors provided the electron emission is sufficient. Tubes that had been used hundreds or thousands of hours, and even tubes that had been rejuvenated several times tested substantially the same as new tubes.

TABLE I

Average Detection Characteristics of Tube Types

Type	Number tested	v (flat part)	R_g at start of flat part (approximate)	μ	R_p (45 volts)	Highest undistorted frequency
201-A	7	0.47	150,000	8	14,000	3,500
200-A	3	0.47	50,000	20	30,000	11,500
240	2	0.47	150,000	30	150,000	3,500
199	6	0.50	125,000	6	17,000	4,250
120	6	0.45	125,000	3	8,000	4,250
171-A	4	0.28	200,000	3	2,500	2,600
112-A	5	0.26	150,000	8	9,000	3,500
226	7	0.29	150,000	8	9,000	3,500
227	6	0.23	50,000	8	10,000	11,500
12	2	0.26	50,000	6	17,000	11,500
102-D	3	0.27	100,000	30	90,000	5,300
310	1	0.45	150,000	8	10,000	3,500

Note: Values for v are average over entire flat portion for tubes listed as tested.
All tubes made by R.C.A. except 102-D which is a Western Electric tube.
Highest undistorted frequency is highest frequency reproduced at least 70 per cent as well as low notes with grid condenser plus tube input capacity of 300 $\mu\mu f$.

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Comparison of Tube Types. The results of an extensive series of tests are presented in Table I. The figures given for each tube type represent the average of all tubes of that variety tested. Individual readings practically never differed by more than 15 per cent from the means tabulated for the voltage constant v.

The rectifying ability (i.e. the change of grid potential produced by a given signal) of the different tubes of Table I can be directly compared at the same grid resistance because the grid-filament tube capacity of all types shown is so near the same as to make it common custom to use the same grid-condenser capacity in all cases. When operated at grid resistances that are in the flat part of the characteristic, the rectifying ability is inversely

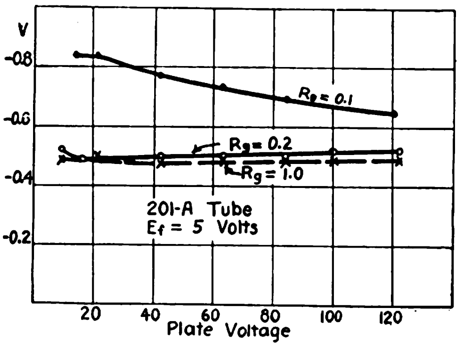

Fig. 5—Effect of Plate Potential on Detector Voltage Constant.

proportional to v, so that the tube with the smallest v (the 227 type) gives the greatest change of grid potential with a given signal, and the tube with the largest v (the 199 type) gives the smallest change of grid potential.

The actual sensitivity of a tube acting as grid leak-condenser detector depends not only upon the rectifying ability of the tube, but also upon the extent to which the change of grid potential is amplified. Thus in 199 tube is not as sensitive a detector as the 201-A type because although both have substantially the same voltage constant, the 201-A type is a better amplifier because of its higher μ and lower plate resistance.

The values of voltage constant v given in Table I for the flat portion of the $v-R_g$ *characteristic apply for all plate voltages above 16 volts up to at least 122 volts (or until the plate current is excessive, as in low* μ *tubes), and for all filament voltages above those giving electron*

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saturation. The value of grid resistance at which the flat part of the $v-R_g$ characteristic begins is lowered very slightly by increasing the plate voltage, and also, at least up to a certain point, by increasing the filament voltage.

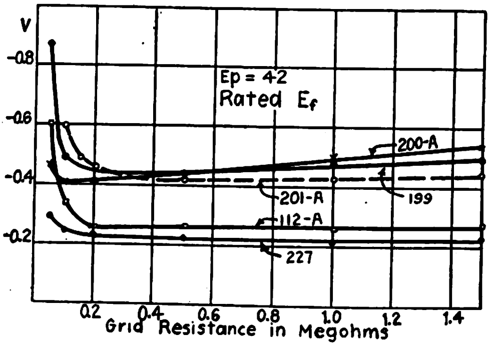

Fig. 6—Detecting Characteristics of Typical Types of Tubes.

A comparison of the rectifying ability of the principal detector tubes now in common use is given in Fig. 6, which includes a typical representative of each type.

The Type 200-A Alkali Vapor Detector. Although this tube contains gas its characteristics are very similar to those of high

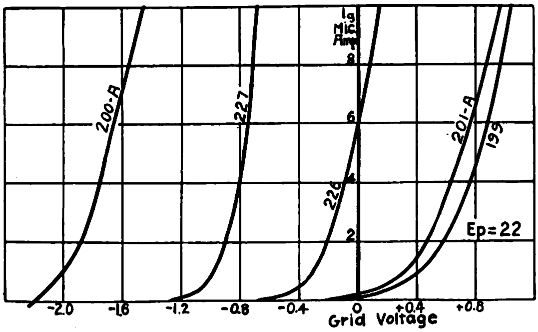

Fig. 7—Grid-Voltage Grid-Current Characteristics of Standard Tubes.

vacuum tubes, as evidenced in Fig. 6. The characteristics of the 200-A tube change a little with variations in plate voltage, but seem to be unaffected by moderate use, or by filament temperatures provided electron saturation is present. The voltage constant of the alkali vapor tubes is constant above a grid resistance of less than 50,000 ohms.

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As a rectifier the alkali vapor tube is almost exactly the same as the 201-A tube, all superiority of the 200-A tube as a detector coming from its higher μ, and from the fact that its wide range of the flat part of the characteristic makes possible high-quality reception without impairment of sensitivity.

The Grid-Voltage Grid-Current Characteristic. The grid-voltage grid-current characteristic of different tube types is similar in general form, but differs in steepness, and in location along the grid-voltage axis, as shown in Fig. 7.

The more sharply the curve bends upward the smaller the voltage constant v and the better the detector. Displacement along the grid-voltage axis is of no importance other than affecting the grid leak required to give a desired operating grid resistance. It will be observed that the tubes with the lowest cathode voltage drop are generally farthest to the left, as one would expect. Decreasing the plate voltage, or increasing the filament voltage, displaces a curve to the left without materially altering its shape. At very negative grid voltages the grid-current of the 200-A tube is seen to go through zero and reverse as a result of ionization.

Discussion

The conclusions of this paper are based on approximately one thousand separate measurements of detector voltage constant v, of which about one third were of an exploratory nature, and are not used in the final results. These measurements were on the average correct to probably five per cent. The results on the same tube taken at different times on the same set-up always agreed within this limit, as did results taken on three different bridge outfits, two of which used the Wagner earth connection while the other did not. If desired, the precision could be greatly improved, but for this study it was preferred to make a large number of measurements of fair accuracy rather than a few very correct ones.

A number of integrity tests were made to check the accuracy of measurements and the correctness of the general theory. These were made by applying a known unmodulated signal to the grid leak-condenser detector when at a known operating point, and measuring the change of plate current. Measured values averaged 5 per cent above the predicted theoretical quantities, with a maximum variation about this mean of approximately 5 per cent.

The grid resistance at which the flat part of the $v-R_g$ characteristic begins is one of the most important properties of a detector

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tube, as it determines the highest audio frequency that the grid leak-condenser detector can reproduce at full sensitivity. The detector voltage constant v increases in value so rapidly as the grid resistance is lowered below the limit of the flat portion that operation much below the bend of the curve is entirely impractical because of the great loss in sensitivity. The end of the flat part of the $v-R_g$ characteristic therefore sets the minimum grid resistance that will give satisfactory sensitivity, and this minimum grid resistance fixes the highest audio frequency that can be reproduced without undue distortion. The last column of Table I gives the highest audio frequency that is reproduced at least 70 per cent as well as the low notes when the operating point is at the grid resistance marking the end of the flat part of the $v-R_g$ characteristic, and the conventional size grid condenser is used.

The points which it is desirable to incorporate in a detector are: (1) a small value of voltage constant v, which gives a large change of grid potential with a given signal; (2) a $v-R_g$ characteristic which is substantially flat, or at a minimum, at grid resistances in order of 100,000 ohms; and (3) a high amplification of the change of grid potential. It is also helpful to have a small inter-electrode capacity in order to permit the use of a small grid condenser. The amplification of a tube is proportional to $\mu/\sqrt{R_p}$ where power output is desired, but is proportional to μ in the case of resistance or impedance-coupled amplifiers.

The selection of a grid leak-condenser detector tube depends upon the circuits involved. When compared on the basis of the audio-frequency power output obtained from a given signal, and disregarding quality, the figure of merit is $(\mu / \sqrt{R_p}) / v$. On this basis the 227 heater type tube is very definitely superior to all other standard receiving tubes. The types 112-A, 226, and 12 are approximately tied for second place, and are followed by a third group made up of types 240, 171-A, and 200-A. The next group includes types 201-A and 199, while the 120 type is the poorest detector of the entire group. If quality of output is taken into account the types 227, 12, and 200-A gain an additional advantage over the rest. When resistance-coupled amplification is used the figure of merit is μ/v, making types 102-D, 240, and 200-A preferred in the order named. It is worth noting that the 112-A tube is very superior to the 201-A type as a detector when both are at the same plate voltage and filament power.

A comparison of tube types 201-A, 240, and the gaseous 200-A

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indicates that a tube similar to the 201-A or the 240 but with $\mu=20$ would have almost exactly the same voltage constant v as the 200-A. The alkali vapor in the 200-A tube would therefore seem to be of no advantage except insofar as it extends the flat portion of the $v-R_g$ characteristic into low grid resistances.

The best plate voltage to use on a detector is ordinarily the highest available, subject to the limitation set by the allowable plate current. At least 45 volts should always be used, and generally 67 to 90 is better. A high plate voltage lowers the plate resistance and does not affect the voltage constant at a given grid resistance. Only in the case of resistance or impedance coupling is the plate voltage unimportant.

The filament voltage is relatively unimportant provided it produces sufficient electron emission to give saturation.

The grid resistance at which the detector operates is determined by the grid-leak resistance and the potential of the grid return lead. The operating point can be conveniently controlled by the resistance of the grid leak.

After considering the detector characteristics of tubes now available one naturally wonders just what design and construction features give a sensitive detector tube. While the results of this paper do not answer this question, analysis of the data does indicate that voltage drop in the filament,[4] type of filament (oxide or thoriated), power rating of tube, and changes in μ are not the most important elements. An investigation is now being made into this and other features of detectors, and it is hoped that definite conclusions can be reported at some future date.

The authors wish to acknowledge their appreciation of the assistance rendered by Messrs. R. Whittern and E. R. Adams in supplying the vacuum tubes used in this study.

[4] The general tendency is for oxide coated filament detector tubes to be superior to other types, but there is a wide variation in oxide coated filaments, some of which (as a J tube tested) have voltage constants of over 0.40 volt.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

FILTERING ANTENNAS AND FILTER-VALVE CIRCUITS*

By
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(Chief Engineer, Polish Marconi Company, Warsaw, Poland)

***Summary**—Some methods of coupling together many circuits or antennas giving them simultaneous excitation from the same source of energy are described. The purpose of such arrangements is the construction of practical filter circuits (filtering antennas) giving square-topped resonance curves with good efficiency. Some interesting phenomena with coupled antennas are described.*

THE importance of square-topped resonance curves in the field of wire telephony was recognized many years ago. Many useful circuits have been developed and constructed by Campbell and others. The importance of square-topped resonance curves for radio work has often been discussed and described. The Marconi Wireless Telegraph Company, Ltd., has used some filters of this kind in their commercial receivers;[1] other companies have probably made use of them also.

In radiotelephony, especially with broadcast receivers, "square-topping" did not attract much attention, since more pronounced and more obvious problems were considered more important at the moment.

F. K. Vreeland[2] has reported some filter circuits in his design for receivers, giving square-topped resonance curves and pointing out their usefulness for distortionless reception of a modulated wave. The author has been interested in this problem for a long time, and has also developed some interesting circuits and antenna arrangements[3] which will be described herein. As the importance of such curves is recognized, the author will limit this paper to the special arrangements developed by him. The filter circuits which were developed consist of resonant circuits coupled together and simultaneously excited by the same source of energy

* Dewey decimal classification: R386. Original manuscript received by the Institute, December 28, 1927. Revised manuscript received June 12, 1928.

[1] See Marconi Pamphlet No. 230/2.

[2] F. K. Vreeland, "On the Distortionless Reception of a Modulated Wave and Its Relation to Selectivity," Proc. I.R.E., **16**, 255; March, 1928.

[3] British Patents 211.151 (1923); 271414 (1926) French Patent. 576.785 (1923). U. S. A. Patent Serial No. 690.120.

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(electromagnetic wave, anode circuit of a valve, etc.) The output connections are attached to one of the parallel circuits. The performance of this arrangement can be determined from the description given below.

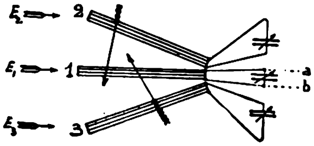

Fig. 1

If we have n antennas mutually coupled (see Figs. 1, 2, and 3) or resonant circuits which receive energy simultaneously from a high-frequency source (i.e., an electromagnetic wave), the differential equations of these circuits can be easily written and the solution indicates that in every circuit both forced oscillations of the same frequency as the source of energy and free oscillations will be generated. These free oscillations represent a number of different frequencies. As a result of the damping of the circuits, all free oscillations disappear after a short time, while the forced oscillations remain.

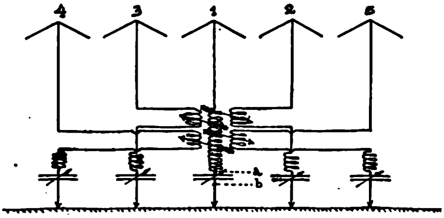

Fig. 2

It can be shown that the time of the disappearance of the free oscillations or the transitory condition of such circuits is shorter than that of ordinary series band filters.[4] The forced oscillations can easily be calculated by well-known methods. We shall first examine two parallel circuits or antennas excited simultaneously:

(1) for a constant incoming wavelength and variable tuning of the circuits, (2) for a variable incoming wavelength and constant tuning of the circuits.

Two Parallel Circuits or Antennas Coupled Together

It can be shown that if we have two circuits or antennas

[4] Campbell, Wagner.

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coupled together the absorbed energies can be calculated by means of the following equations:

$$W_1=(i_1)^2_{eff}\times R_1=\frac{R_1}{2}\times\frac{(E_2y-E_1Z_2)^2+E_1^2R_2^2}{[Z_1Z_2-(R_1R_2+y^2)]^2+(R_1Z_2+R_2Z_1)^2}$$

$$W_2=(i_2)^2_{eff}\times R_2=\frac{R_2}{2}\times\frac{(E_1y-E_2Z_1)^2+E_2^2E_1^2}{[Z_1Z_2-(R_1R_2+y^2)]^2+(R_1Z_2+R_2Z_1)^2}$$

where

$$y=M\omega$$

$$Z_1=\left(L_1\omega-\frac{1}{C_1\omega}\right)\quad Z_2=\left(L_2\omega-\frac{1}{C_2\omega}\right).$$

For a constant frequency of the electromagnetic wave and constant tuning of the first circuit, the effect of the degree of

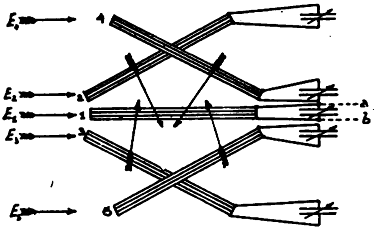

Fig. 3

coupling is very interesting when the tuning of the second circuit is made variable.

In Fig. 4 curves for two circuits are given. Power absorbed is plotted as a function of the tuning of the second circuit. The first circuit is tuned to the incoming wavelength, then it is coupled with the second circuit, which is then tuned by means of a tuning condenser. Near the resonance point of the second circuit we have a minimum of current in the first circuit, then a maximum, and the asymptotical currents decrease. The maxima and minima depend on the coupling. This phenomenon is clearly shown in Fig. 4. In the second circuit we have ordinary resonance curves, although the apparent resistance is higher.

Judging by these curves the current at its maximum in the first circuit coupled with the second circuit is greater than in the first circuit alone, not coupled with the second circuit. The mutual coupling of the two circuits reduces the apparent resistance of the first circuit. It is obvious that in this case the first

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circuit receives energy not only from the electromagnetic wave, but also from the second circuit. From Fig. 4 we can easily calculate the efficiency of this arrangement, which can be exceedingly great (90 per cent or more).

In series filters the efficiency must be much less because the efficiencies of different chain members of the filter must be mul-

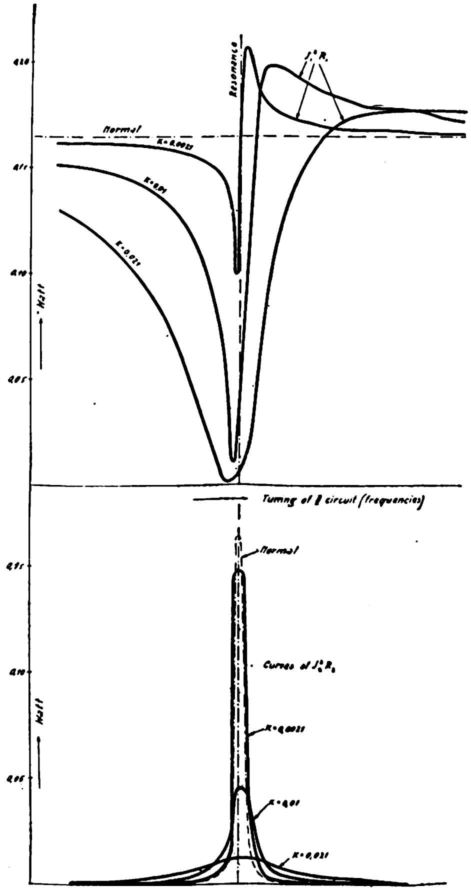

Fig. 4

tiplied in order to obtain the efficiency of the whole system. The over-all efficiency of a series filter is indeed very low (about 50 per cent), depending on the number of sections. It can also be shown that for several parallel circuits the same effects can be obtained, and that the efficiency of such filters is greater than

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the efficiency of series filters. If the effective resistances of the two coupled circuits are different, very sharp resonance curves are obtained. The relation between minimum and maximum can be very great.

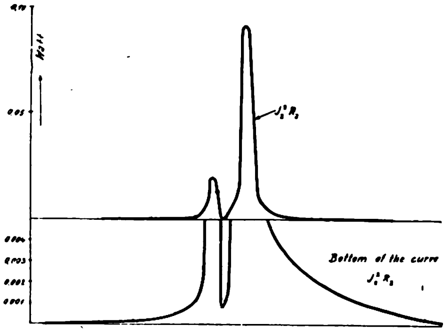

Fig. 5a

For all curves mentioned above it was assumed that the electromotive forces were equal. If they are much different we observe the ordinary effect of the two coupled circuits with an electromotive force in one of them. The more the electromotive force appears in one of the circuits the more the minima and maxima are smoothed out.

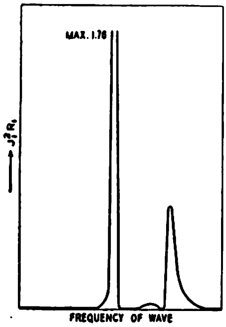

Fig. 5b

The above-mentioned effects could be obtained by exact calculation of the electrical values of the two circuits. It is obvious from the above-mentioned figure that the effective resistances of the two circuits must have the correct relation to the coupling degree of the two circuits.

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$$\omega M \leqq R_1;\ \omega M \leqq R_2;\ R_1 \leqq R_2$$

The position of minimum and maximum can be changed if we shift the coupling coil through 180 deg. or if we change the phase of the electromotive force by 180 deg. In this case we have first maximum and then minimum, if we go from a lower to a greater frequency.

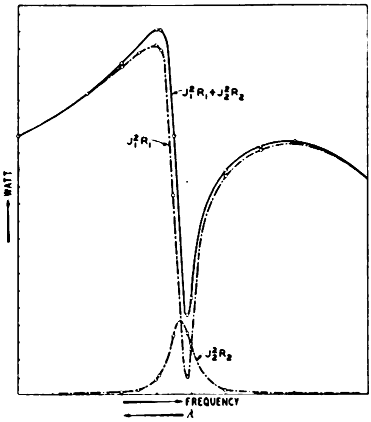

Fig. 5c

Constant Tuning of Circuit I and Circuit II with the Wavelength of the Electromagnetic Wave Variable

We take the same circuits as described in Fig. 4, and change the frequency of the oscillations. Again we obtain deformed resonance curves. The curve $J_1^2R_1$ has a flat minimum and a maximum. The curve $J_2^2R_2$ has two maxima and one minimum. In Fig. 5 the curves of $J_2^2R_2$ are represented. In Fig. 5b the curve of $J_1^2R_1$ is shown when $R_2 << R_1$. In Fig. 5c the curves of $J_1^2R_1$ and $J_2^2R_2$ are shown for circuits with $R_1 = R_2$ but with different values of C and L.

Three and More Circuits in Parallel

The calculation of mathematical figures for three and more circuits or antennas in parallel is very difficult as the corresponding figures are complicated. In order to facilitate this work we put some of the coefficients of mutual induction equal to zero. Such

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a simplification of the problem is permissible because we can always construct an antenna or a circuit so that some of the coefficients of mutual inductance will equal zero. It is very difficult, often impossible, to construct three or more circuits with the same coefficients of mutual inductance.

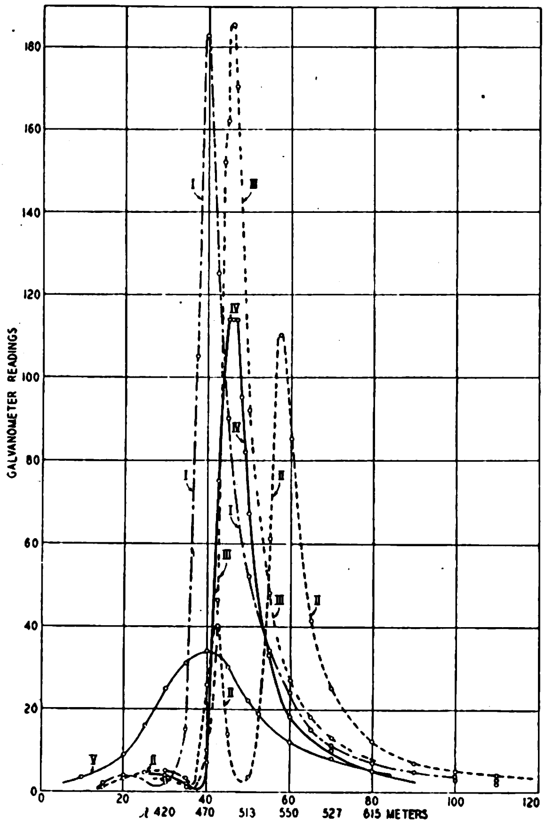

Fig. 6

Let us take the circuits as shown in Fig. 1; we can put $M_{23}=0$; that is, the coupling between the second and third circuits equals zero. We can obtain the same condition for more than three circuits. In this case the corresponding mathematical figures are deduced more readily.

If we have three circuits (Fig. 1), we can put $M_{23}=M_{32}=0$,

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and we thus obtain the corresponding figure for absorbed energy:

$$W_1 = R_1 I_1^2{}_{eff} = \frac{1}{2} \times$$

$$\times \frac{R_1 \times \left\{ \left[E_1 - E_2 \frac{Z_2 y_2}{Z_2^2 + R_2^2} - E_3 \frac{Z_3 y_3}{Z_3^2 + R_3^2} \right]^2 + \left[E_2 \frac{y_2 R_2}{Z_2^2 + R_2^2} + E_3 \frac{y_3 R_3}{Z_3^2 + R_3^2} \right]^2 \right\}}{\left[R_1 + \frac{y_2^2 R_2}{Z_2^2 + R_2^2} + \frac{y_3^2 R_3}{Z_3^2 + R_3^2} \right]^2 + \left[-Z_1 + \frac{Z_2 y_2^2}{Z_2^2 + R_2^2} + \frac{Z_3 y_3^2}{Z_3^2 + R_3^2} \right]^2}$$

where

$$\begin{cases} Z_1 = \left(L_1 \omega - \frac{1}{C_1 \omega} \right) \quad Z_2 = \left(L_2 \omega - \frac{1}{C_2 \omega} \right) \quad Z_3 = \left(L_3 \omega - \frac{1}{C_3 \omega} \right) \\ y_2 = M_{12} \omega, \quad y_3 = M_{13} \omega \end{cases}$$

We shall now study the change of currents: (1) for a given wavelength of the electromagnetic wave, and (2) for a given tuning of the three circuits as a function of a variable wavelength of the received wave. For a constant frequency of the wave with

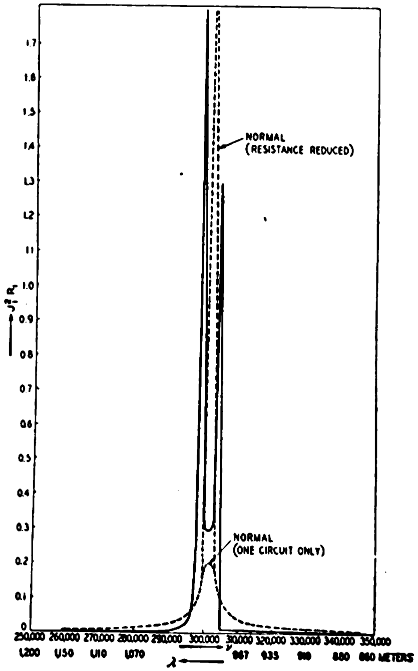

Fig. 7

the first and second circuits tuned to maximum current in the first circuit and variable tuning in the third circuit, a phenomenon similar to that of two circuits in parallel is obtained; there is a minimum, a maximum, etc.

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For a Variable Frequency of Wave and Constant Tuning of the Three Circuits

Here we attain very interesting curves. Let us discuss a practical case (Fig. 6). We have frame aerials (closed circuits) as indicated in Fig. 1. The first frame was connected to a galvanometer through a detector. The second and third frames were removed and Curve V was taken for the first frame by changing the heterodyne frequency; that is an ordinary resonance curve. After this the second frame was coupled with the first and adjusted in order to give the maximum of current in the

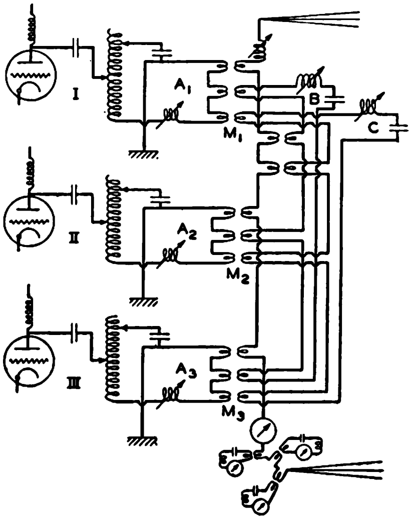

Fig. 8

first frame for $\lambda = 470$ meters. Then the third frame was coupled and adjusted to obtain still more current in the first frame. This done the curve for the current of the first frame was checked (Curve I). Then the second frame was slightly detuned by increasing its capacity about three per cent and the third frame detuned by diminishing its capacity about three per cent. The corresponding curve has two maxima (Curve II). If the detuning of the second and third frames is diminished a setting can thus be obtained where Curve IV appears, and the filtering effect of this arrangement becomes evident. This arrangement received with uniform intensity a band of frequencies approximately 20,000 cycles wide. The frequencies lying outside this band are

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strongly damped; Curve IV on either side of the drawing goes below Curve V. If the detuning of the second and third frames is further diminished, Curve III is obtained, which is similar to a normal resonance curve. It must be pointed out that all curves have a pronounced minimum. It is obvious that with five and more circuits still better results can be obtained.

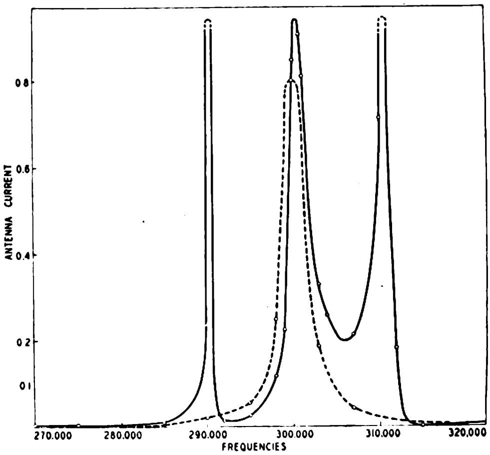

Fig. 9

Taking three frames, with the second and third having resistances equal to zero by reaction, a resonance curve is obtained, as in Fig. 7. That is an inverted resonance curve.

The same arrangement can be used for a multiplex transmitting system in wireless telephony and telegraphy.

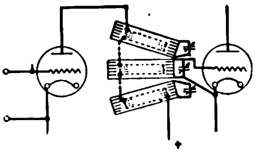

Fig. 10

Fig. 8 shows a system of multiplex transmission. A radiating antenna is coupled with two closed circuits B and C, influenced by a transmitter's intermediate circuit A_1 and coupling M_1. The intermediate circuit influences at the same time the two closed circuits B and C; transmitter I excites the antenna by the intermediate circuit A_1 and allows radiation of energy on a wave λ_1.

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Transmitter II acts by the intermediate circuit A_2 and coupling M_2 upon the same aerial and the same coupled circuits B and C. It is possible in this way to transmit on the same antenna a second wave λ_2. Transmitter III acts by the intermediate circuit A_3 and coupling M_3 upon the antenna and the same closed circuits B and C, allowing transmission of a third wave λ_3. In

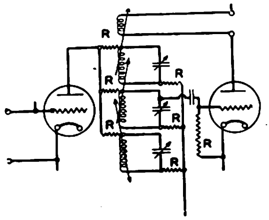

Fig. 11

such a manner it is possible to transmit simultaneously many waves from one radiating antenna. If in general there are $(n-1)$ closed circuits coupled with the antenna, n messages can be transmitted simultaneously.

With correct tuning of the coupled circuits a band of frequencies amounting approximately to 10,000 cycles will be radiated in a square-topped resonance curve. (Fig. 8) In this case we

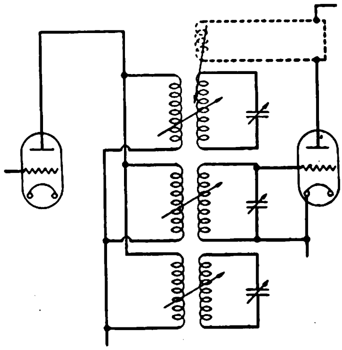

Fig. 12

must have some independent drive circuits connected to the grids of amplifying valves in such a way that the modulated wave will be transmitted by the antenna system without distortion. If multiplex transmission is to be obtained by the arrangement shown in Fig. 8 many drivers tuned to the respective wavelengths of the antenna system are required. In this case the

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characteristic curve of the antenna will be as indicated in Fig. 9. The curve shown by dotted lines is the resonance curve of the antenna with circuits *B* and *C* removed.

It is possible to use one driver circuit and one amplifying system together with the antenna system described. In case of modulation of the wave by many frequencies (acoustic or supersonic) so that the resulting modulation waves are in resonance

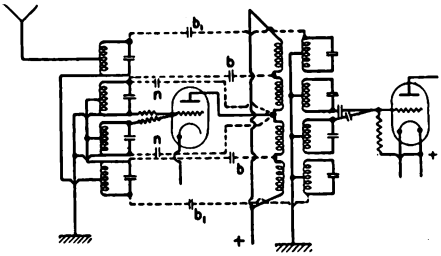

Fig. 13

with the antenna system, a multiplex transmission can be obtained very simply.

The same may be said of an antenna system used for reception, but in this case we must have many parallel open aerials or many parallel frame aerials; otherwise simultaneous excitation of all circuits by incoming waves cannot be had. This applies to intervalve filters also. For instance, if we wish to obtain an intervalve

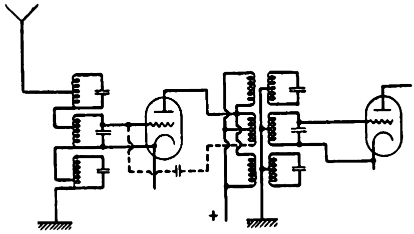

Fig. 14

filter with square-topped resonance curve or some other curve suitable for good transmission or reception, we must construct according to Figs. 10, 11, 12, 13, and 14. On these drawings *R* indicates non-inductive resistances. Intervalve filters of this kind can be constructed also to obtain resonance curves with many peaks like those of Fig. 9. This arrangement can be effected either by means of variable condensers or inductances

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tuned to different wavelengths, or with fixed condensers or inductances, for instance, in superheterodyne receivers.

The same types of intervalve transformers can be utilized for amplification of a number of waves for multiplex transmission, amplification, or reception. The regenerative effect may be used and reaction coils applied (Figs. 11 and 12). These coils may be coupled with one or with all circuits, or special reaction valves and reaction circuits not connected directly to the whole system may be applied. The various possible schemes are indicated in the drawings, which are self-explanatory.

The curves obtained experimentally have been found similar to those reported in the paper by Mr. F. K. Vreeland,[2] and Figs. 6, 7, and 9 of this paper.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

Discussions on THE RECEIVING SYSTEM FOR LONG-WAVE TRANSATLANTIC RADIO TELEGRAPHY*

Austin Bailey, S. W. Dean, and W. T. Wintringham

(At meeting of Washington, D. C., Section of the Institute, October 11, 1928)

F. P. Guthrie[1]: The most interesting points are always brought out in the discussion after the paper. I have a couple of questions I want to ask. You state you get a great improvement by going up into Maine, but there is a lot of difference between Maine and the North Pole. Could you go farther north and get greater improvement?

Austin Bailey[2]: Maybe you could. I think it is an economic question. Wire circuits cost money and power costs money, and it is a question of where you can get the best communication circuit for the least cost. If you figure you can obtain a good and more economical circuit by moving farther north, then move farther north. The transatlantic radio circuit is young yet. It is only two years old. And some further improvements may be obtained by just such means as that.

F. P. Guthrie: Does that mean that our Canadian friends have a great advantage over us always?

Austin Bailey: I think it does mean some advantage for long waves. If they are going to communicate with England, they are nearer to England, and farther away from primary sources of static.

A Member: Couldn't you get far enough north that the aurora borealis would be serious, or just what does that do to the transatlantic signals?

Austin Bailey: I have heard it twice in Cupar, Scotland, and it is a very interesting thing to hear. It gives a sort of a hissing noise. I imagine a good many people besides myself have heard it. It is a source of interference, and it may be serious, but as yet we have no data at all to indicate how serious.

A Member: During time of aurora borealis there is a very great decrease in the field strength of the signals—greatest

* Proc. I.R.E., **16**, December, 1928.
[1] Radio Corporation of America, Washington, D. C.
[2] Department of Development and Research, American Telephone and Telegraph Company, New York City.

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especially in the ordinary broadcast wavelengths. I have not really observed it in the longer wavelengths.

Austin Bailey: First of all, it is probable, I believe, that the aurora borealis and the variations of the magnetic field of the earth are due to a common cause. The variation in signal field strength observed in the broadcast frequency range, and much more so at higher frequencies, is probably due to magnetic field variation or to the same thing that causes both of the other things. I cannot answer what it is due to, but the fact remains that there is some correlation between these things. On the very high frequencies increase in disturbance of the magnetic field is, in general, accompanied by very low signal fields. At such times on long wavelengths you will get an increase in the daylight field strength and a decrease in the night-time field strength.

August Hund[3]**:** The speaker has shown us on the screen the transmission formulas due to Sommerfeld, Austin and Cohen, and Fuller, and one worked out by the Bell System and mentioned that each formula gave about the same results for the frequency band and distance in question. I wonder whether the speaker has ever tried to find out from these formulas what distance should be used for a certain frequency for the best reception at the other end. The formula lends itself very readily to such an experiment, because it is only necessary, for instance, to differentiate with respect to the frequency and solve for a maximum, or to differentiate with respect to the distance and solve for a maximum.

Austin Bailey: There is, of course, more to the problem of choosing the best frequency for a given distance than is indicated by Dr. Hund's thought. The best reception depends also upon the transmitting antenna efficiency and the static at the receiving station as shown by Figs. 1 and 2 of the paper. Furthermore, the transmission formula was derived for the particular transatlantic path under consideration, and how well it will work for any other transmission path I do not know.

August Hund: In the course of his lecture the speaker has given us a very ingenious method for getting the characteristic impedance as well as the propagation constant of his wave-antenna. The method to which I have reference measures these quantities by using two known terminating impedances at the

[3] Radio Section, Bureau of Standards, Washington, D. C.

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end of the line. The reason was, if I understood it right, because the usual method, which carries on the measurement for the open-ended and shorted line, would give either unusually small or unusually large values, which would make the measurement very difficult. I just wonder whether it would not be possible to measure the characteristic impedance simply at any point by noting the potential at that point and the current through that point?

Austin Bailey: The only thing you get by doing that is the absolute value of the impedance. You cannot measure the phase angle; therefore, you cannot get the propagation constant.

August Hund: In case of a wave-antenna, we have to adjust at the transmitting end exactly to the value of the characteristic impedance. Now, I just wonder whether, for the different weather conditions, this surge impedance can be considered as constant.

Austin Bailey: As an actual matter of fact, we have made measurements over an extended period of time to determine that. The variations are remarkably small. Surface moisture and the depth to which the ground freezes do not seem to have any large effect on the antenna characteristics. The ground waves and constants are not greatly affected except by slow changes over long periods of time.

August Hund: When Dr. Austin and Dr. Cohen gave us their empirical transmission formula it worked all right for a number of years. The reason for this was that for the first series of their experiments they were working only over comparatively short distances, but as soon as Dr. Austin extended his work over considerable distances he felt the need of changing his formula in order to agree with his measurements. The reason is probably due to the fact that even in the original Sommerfeld formula we find only one exponential term. Although I do not believe in making a formula more complicated than it is already, it would seem that a formula with two exponential terms might give more reasonable values. The reason for this is because when we receive over very, very great distances sometimes a wave may go around the Great Circle one way just as well as the other way. As a matter of fact, such cases have been reported in the literature where the signal was received over each path. (Of course, in this consideration, I am not thinking of the Heaviside layer and its effect.) Now, if we should take two exponential

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terms, one holding for one path and the other for the other path, it might perhaps explain why the intensity over long distances should change when only one term is used, as in the present empirical formula.

Austin Bailey: I think that there have been proposed a few transmission formulas to take into account overland and overwater transmission, but I do not believe that anyone has previously proposed your idea of transmission in two directions.

A. H. Taylor[4]: In working at the R.C.A. station at Chatham, Massachusetts, at night I noticed that the signals from Long Island could be barraged out for only a few moments at a time. I was interested in that because the variation in direction is present at night. These variations seem to be maximum around 200 miles depending somewhat on the frequency. At night here in the winter time, provided with excellent compensation there are instantaneous fluctuations of 90 deg. on New Brunswick direction and are very rapid. I have never made any measurements exactly at the frequency you are working on. Do you know what those variations are? They should be much less on transatlantic stations.

Austin Bailey: If you receive at night you run into this effect. We received at Houlton for these wave-antenna tests only in the early morning and late evening. We could not work at night because of this, and in the daytime the station was in commerical use on traffic, and so the only available times were the little pieces at the beginning of the morning about 4 to 7 A.M. and in the evening from 5 to 7 P.M. Now, as to the magnitude of this variation in the apparent direction, the signals from the mobile transmitter at a distance of about 50 miles frequently varied in direction as much as 20 deg. during the period immediately preceding sunrise. Little advantage can be taken of the directional diagram at night, but fortunately you get a fairly large signal-to-noise ratio at night and you usually can work satisfactorily. It is in the afternoon in the United States when the static comes in that the directivity is essential. You would be absolutely lost if you did not have directivity at this time of day when the signal is decreased due to the intervening shadow wall between the transmitting and receiving stations, but the static is being received mostly over an all-daylight transmission path from the opposite direction from the signal.

[4] Radio Division, Naval Research Laboratory, Bellevue, Anacostia D. C.

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A. H. Taylor: We had a station at Bar Harbor, one at Long Island, one in New Jersey and at Washington. We found a progressive increase in the signal ratio especially as you went along. That is, New Jersey was distinctly better than Washington; Long Island, with inferior receiving apparatus, better than New Jersey; Chatham was distinctly better by a definite amount than New Jersey, and Bar Harbor was above everybody, not only in the signal-noise ratio, but due to the fact that the total signal intensity was tremendous up there, and you could copy from stations on much higher frequencies. During the years of the development of the Radio Corporation, it was the consensus of opinion among all of us engaged in transatlantic telegraphy that for the general New York-Washington location there was no use fooling with anything short of 12,000 meters. It simply did not go during the summer months. Are we to infer from these more recent developments with modern amplifiers, that the opinion ought to be revised? We based that opinion on the signal-noise ratio which should not be affected by amplifiers, and we also had the use of fairly directive receiving equipment. Am I to believe now that it would be possible to drop the wave for transoceanic telegraph and get communication in the summer time?

Austin Bailey: There is one other factor to be taken into consideration, and that is the efficiency of the transmitting antenna. Take that factor into consideration and build an antenna with Rocky Point efficiency, and you will probably get better communication service at higher frequencies even in the summer than you would on the longer wavelengths.

A. H. Taylor: If you could not go up to Maine and received it somewhere between here and New York, could you do it?

Austin Bailey: We do not have any data for locations between here and New York to answer that. On Long Island the higher frequencies are undoubtedly preferable. Of course, a directive system located as far south as New York is at great disadvantage, because a large number of storms occur in front of it, and there is no hope during these periods. At Houlton there are about 12 thunderstorms a year on the average.

A. H. Taylor: All downcoming waves have three components. In connection with the higher frequencies we have been doing some work with these components. We have fallen into the habit in our laboratory of referring to three components as the vertical, horizontal, and longitudinal thrust. The idea is generally appli-

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cable to all frequencies. In order to distinguish the two horizontal components I suggest the terms "horizontal component" and "longitudinal thrust."

Austin Bailey: That is a very good suggestion. The two components, of course, in this paper, are quite intimately connected with the primary plane polarized transmitted wave, and that is the reason that we have used the two primary terms, vertical and horizontal, in connection with our work rather than vertical and longitudinal.

A. H. Taylor: By the way, do you find your night static less centered in direction than the daylight static?

Austin Bailey: Yes, more scattered, and probably partly due to the night-time directional variations already mentioned.

F. P. Guthrie: Are there any essential differences between a good antenna for telephony and a good antenna for telegraphy?

Austin Bailey: One is that the Beverage-Rice-Kellog compensation method is perfectly satisfactory for telegraphy. They are interested in balancing out disturbances at one frequency only. In telephony we have to balance out a band 3000 cycles wide, so we should have for telephony a balance which is invariable with frequency. We can turn these blind spots around against static and still maintain them over the whole telephone band range and not just at one frequency. A satisfactory antenna for telegraph receiving may not be so satisfactory for telephone receiving but not the other way around.

A. H. Taylor: Do you think that short-wave telephone transmission can ever be as good as long-wave telephone transmission?

Austin Bailey: It depends for one thing upon the power you use. If you use the same amount of power on short waves as on long waves, I should think you would get about equal reliability. Long and short waves are vulnerable from different causes. You can use small power on short waves and get away with it part of the time.

(At New York meeting of the Institute, November 7, 1928.)

Alfred N. Goldsmith[1]: This paper is a clear indication of the evolution which follows the first exciting, almost "freak" reception, (if we may use the term for reception accomplished only

[1] Chief Broadcast Engineer, Radio Corporation of America, New York City.

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under especially favorable conditions,) to a steady and reliable, thoroughly planned, and capably engineered service. To listen to a paper of this sort is, therefore, not merely to receive technical information, but also to have traced before one the evolution of a portion of a new art.

In the midst of an extremely interesting discussion of a complicated subject, I hope that no one will go away with the impression that we have learned so much about the possibility of "improving" broadcasting by eliminating one side-band and suppressing the carrier that this method of transmission will be seriously proposed for ordinary broadcasting. One gets a disquieting picture of a few hundred thousand independent local oscillators working in New York City, for example, on frequencies selected by the casual listener in his operation of the receiver. This is definitely not a happy picture. The incorrect selection of the re-inserted carrier frequency in broadcasting would result in sounds even less harmonious than those which are sometimes experienced by the listener utilizing the more conventional type of receiving set.

So that it is to be hoped that optimism will not lead to the adoption, at least immediately, in broadcasting of these ingenious methods which are so thoroughly applicable in the hands of skilled operators working on a single frequency in a transatlantic telephone receiving station.

K. S. Van Dyke[2]: Will the authors mind stating what the order of magnitude of their ground connection impedance is, particularly with relation to the characteristic impedance of the antennas?

W. T. Wintringham[3]: Of the four antennas at Houlton, the characteristic impedances roughly average about 420 to 425 ohms. The lowest values aren't much below 400, and the highest values not much over 450.

A poor ground, for example the one that was pointed out in one of the slides as being on the top of a rocky hill, measures about 40 ohms, mostly negative reactance, whereas a good ground will measure possibly 1 or 2 ohms resistance, with between 1 and 2 ohms positive reactance.

[2] Department of Physics, Wesleyan University, Middletown, Connecticut.

[3] Department of Development and Research, American Telephone and Telegraph Company, New York City.

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Haraden Pratt[4]: In these days of high-frequency technique, which is absorbing everyone's interest, it is refreshing to see such a learned discourse on the older art of low-frequency radio transmission, particularly in the telephone field, where the conditions are a little more exacting from the standpoint of the customer of the service. It is inconvenient for a telephone subscriber to be interrupted and repeat what he has to say on account of interference, and so forth, which the user of the telegraph service isn't so much aware of—a few repetitions can occur and he doesn't know about them—which fact has probably pushed the development of this receiving problem a little bit further than has been the case in the telegraph field.

The signal-to-noise ratio study has interested me particularly on account of the long statistical work which no doubt was necessary over a period of years in order to arrive at a conclusive result, and it would appear from the paper that the workers on the system are particularly fortunate in finding a combination which gave a suitable frequency from the signal-to-noise ratio standpoint, and also was within financial limits in the building of the receiving antenna system, and, fortunately, fell within the requirements from the standpoint of band width of the transmitting set.

The favorable location of receiving points in the northern part of the country, as Mr. Bailey mentioned, has been observed before, and it might be interesting to recount an experience that occurred early in the World War, when radio communications were being relied upon in case of cable interruption for military communications with Europe. It had been observed that receiving conditions in the northern part of the United States were favorable, more favorable than in the southern part, and therefore an observer was sent to Newfoundland early in 1918 to see what he could do up there in receiving the low-frequency signals from European stations, which at that time were not high in power, and which were not able to put good signals into the United States in the summertime.

That man, after he arrived at Newfoundland, found that a mistake had been made in the instructions given in regard to the equipment which he was to use, and when he unpacked his apparatus found a receiving set that had only a limited range and prevented his receiving the European stations on their usual

[4] Chief Engineer, Mackay Radio and Telegraph Company, New York City.

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frequencies. But transmitting sets in those days radiated harmonics, and during the summer of 1918 this observer sent us copy from Newfoundland of signals from some of the European stations received on their harmonics, which were sent at times when the same signals could not be readily copied at the usual pointsin this country. Of course, Newfoundland had the further advant age of being much closer to the transmitting point.

The diagrams of the geological formation of the ground and the matter of ground resistance being so important I think is particularly interesting, as it shows the trend of the times in the matter of correlating the radio engineer's work with that of the geologist, the physicist, and the meteorologist, and I think it brings out a great many very interesting points in that regard.

F. H. Murray[3]: The use of Raleigh's solution of the problem of random phases in the problem under discussion leads to rather laborious computations; a shorter approximate solution is derived here. Comparison shows a close numerical agreement in some special cases.

From mathematical investigations it is known that the probability that the sum $\sum a_i \cos(V_i t+\delta_i)$ will exceed any fixed number x, the numbers V_i being incommensurable, is the same as the probability that $\sum a_i \cos\theta_i$ is greater than x, where all values of any θ_i are equally likely and these angles are independent variables. In the Dirichlet integral

$$\frac{1}{2\pi}\int_{-\infty}^{\infty}\frac{\epsilon^{i\omega g}}{i\omega}d\omega=\begin{cases}0, & g<0\\ \frac{1}{2}, & g=0\\ 1, & g>0\end{cases}$$

let $g=\sum_{i=1}^{n} a_i \cos\theta_i - x$, then the required probability is

$$P=\left(\frac{1}{2\pi}\right)^{n+1}\int_0^{2\pi}d\theta_1\int_0^{2\pi}d\theta_2\cdots\int_0^{2\pi}d\theta_n$$
$$\int_{-\infty}^{\infty}\frac{1}{i\omega}\epsilon^{i\omega\left[\sum_r a_r\cos\theta_r - x\right]}d\omega$$

Interchanging the order of integration and letting $a_i=a$, we have

$$P=\frac{1}{2\pi}\int_{-\infty}^{\infty}\frac{1}{i\omega}\epsilon^{-i\omega x}J_0(a\omega)^n d\omega$$

where J_0 is the Bessel function of order 0.

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If N is the total number of possible sources of terms $a \cos \theta_i$, p the probability that any one source will contribute its term, and the same for all, the probability that enough terms are present and have the proper phases to make $\sum a_i \cos \theta_i > x$ is equal to

$$\overline{P}(x) = \sum_{n=0}^{N} \binom{N}{n} p^n (1-p)^{N-n} \cdot \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{1}{i\omega} \epsilon^{-i\omega x} J_0(a\omega)^n d\omega$$

$$= \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{\epsilon^{-i\omega x}}{i\omega} [1 - p(1 - J_0)]^N d\omega.$$

To find an approximate value for $\overline{P}(x)$, observe that

$$[(1-p) + pJ_0]^N$$

has its maximum value (for real ω) at $\omega = 0$, and decreases rapidly for large $|\omega|$. Hence only the values of ω near $\omega = 0$ are of importance in the integral $\overline{P}(x)$. Writing

$$\log [1 - p(1 - J_0)] = -p(1 - J_0) - \frac{p^2}{2}(1 - J_0)^2 - \frac{p^3}{3}(1 - J_0)^3 \cdots$$

$$1 - J_0(a\omega) = \frac{(a\omega)^2}{4} - \frac{a^4\omega^4}{64} + \frac{a^6\omega^6}{2304} - \frac{a^8\omega^8}{2^8(4!)^2} \cdots$$

we have finally, for $a = 1$, $Np = A$,

$$N \log [1 - p(1 - J_0)] = -\frac{A}{4}\omega^2 + Aa_1\omega^4 + Aa_2\omega^6 Aa_3\omega^8 + \cdots$$

$$a_1 = \frac{1}{2^6} - \frac{p}{2^5},$$

$$a_2 = \frac{-1}{9 \cdot 2^8} + \frac{p}{2^8} - \frac{p^2}{3 \cdot 2^6}$$

$$a_3 = \left[\frac{1}{(4!)^2} - \frac{17p}{288} + \frac{p^2(1-p)}{4}\right] \cdot \frac{1}{2^8}$$

Consequently

$$\epsilon^{N \log [1 - p(1 - J_0)]}$$

$$= \epsilon^{-A\omega^2/4} \left[1 + Aa_1\omega^4 + Aa_2\omega^6 + \left(Aa_3 + \frac{A^2 a_1^2}{2}\right)\omega^8\right] + \cdots$$

If $z = i\omega$, the required probability becomes

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$$\overline{P}(x)=\frac{1}{2\pi}\int_{-\infty}^{\infty}\frac{\epsilon^{-zx}}{z}\epsilon^{(Az^2/4)}d\omega$$

$$+\frac{1}{2\pi}\int_{-\infty}^{\infty}\epsilon^{-zx+(Az^2/4)}\left[Aa_1z^3-Aa_2z^5+\left(Aa_3+\frac{A^2a_1^2}{2}\right)z^7\right]d\omega.$$

If $B=A/2$, $erf\ (x)=2/\sqrt{\pi}\int_0^x \epsilon^{-t^2}dt$, this becomes

$$\overline{P}(x)=\frac{1}{2}\left[1-erf\left(\frac{x}{\sqrt{A}}\right)\right]-\frac{\epsilon^{-x^2/A}}{\sqrt{\pi A}}\left[Aa_1\left(\frac{3x}{B^2}-\frac{x^3}{B^3}\right)\right.$$

$$-Aa_2\left(\frac{-15x}{B^3}+\frac{10x^3}{B^4}-\frac{x^5}{B^5}\right)$$

$$\left.+\left(Aa_3+\frac{A^2a_1^2}{2}\right)\left(\frac{105x}{B^4}-\frac{105x^3}{B^5}+\frac{21x^5}{B^6}-\frac{x^7}{B^7}\right)\right]+\cdots$$

In the case $N=100$, $p=0.15$, the first term of this sum gives the correct value within a few per cent. if $x=5$.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

Discussion on
DETECTION WITH THE FOUR-ELECTRODE TUBE*

(J. R. Nelson)

F. B. Llewellyn[1]**:** The relations given in Mr. Nelson's valuable paper have suggested several interesting points in connection with detectors in general and screen-grid detectors in particular. Equation (23) in Mr. Nelson's paper,[2] or (31) and (32) in my own paper, give the expression for detected current. This expression for the case where two sine waves of voltage

$$A \cos ht + B \cos kt$$

are applied to the grid gives for the component of plate current of frequency $(h-k)$ the following expression:

$$i_{(h-k)} = \frac{1}{2}\left(\frac{r_p}{r_p + z_{(h-k)}}\right)$$

$$\left\{\frac{\partial g_m}{\partial E_g} - \left(\frac{\mu z_h}{r_p + z_h} + \frac{\mu \bar{z}_k}{r_p + \bar{z}_k}\right)\frac{\partial g_m}{\partial E_p} - \frac{g_m{}^2 z_h \bar{z}_k}{(r_p + z_h)(r_p + \bar{z}_k)}\,\frac{\partial r_p}{\partial E_p}\right\} AB\cos(h-k)t.$$

It is the nearly invariable practice in detector design to make $z_h = z_k = 0$. In that case the above expression reduces to the simple relation,

$$i_{(h-k)} = \frac{1}{2}\left(\frac{r_p}{r_p + z_{(h-k)}}\right)\frac{\partial g_m}{\partial E_g} AB \cos (h-k)t.$$

Moreover, since the static characteristic for $I_p - E_g$ with a fixed value of E_p may be represented over a considerable range by an expression of the form

$$I_p = KE_g{}^n$$

we have

$$\frac{\partial g_m}{\partial E_g} = \frac{n-1}{nI_p} g_m{}^2$$

or, when $n=2$

$$\frac{\partial g_m}{\partial E_g} = \frac{g_m{}^2}{2I_p}$$

* Proc. I. R. E., **16**, 822; June, 1928.

[1] Bell Telephone Laboratories, New York City.

[2] F. B. Llewellyn, "Operation of Thermionic Vacuum Tube Circuits," *Bell System Tech. Jour.*, July, 1926.

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which is convenient for rapid approximations. When a more accurate evaluation is desired, the dynamic method suggested by Mr. Nelson is a simple and reliable one provided small values of input a.c. voltage only are employed.

Since, as shown above, detection depends upon the coefficient $\frac{\partial g_m}{\partial E_g}$ which is equivalent to $\frac{\partial}{\partial E_g}\left(\frac{\mu}{r_p}\right)$, and since for the screen-grid tube μ varies considerably in the operating region, it is evident that detection with that type of tube is accomplished primarily by μ variation. This is in striking contrast to the operation of the ordinary three-element tube where μ is fairly constant, and the greater portion of the detection arises from variation of r_p.

J. R. Nelson[3]: Mr. Llewellyn points out that in using the dynamic method suggested in my report only small input a.c. voltages should be used. This is in general true, and the more sensitive the detector the smaller the a.c. input voltage should be. A screen-grid tube, however, is an exception to the general case as large a.c. input voltages may be applied to the grid of the tube and values of $\partial g_m/\partial e_g$ found will check the values obtained with the smaller a.c. input voltages. This is due to the fact as pointed out in the paper that the variation of G_m with regard to E_g is practically constant in the range studied.

The method suggested by Mr. Llewellyn is a very convenient one for comparing different tubes as bias detectors or finding the best operating point of a tube. The values of G_m and I_p at the operating point are easily found.

Mr. Llewellyn in a letter to the author pointed out an error in (22) and following equations so that $_{(h-k)}a_2$ in (23) to (29) should read

$$_{h-k}a_2=\frac{r_p}{2}\,\frac{1}{r_p+z_{h-k}}\left\{\frac{\partial g_m}{\partial e_g}-\left[\frac{\mu\bar{z}_k}{r_p+\bar{z}_k}+\frac{\mu z_h}{r_p+z_h}\right]\frac{\partial g_m}{\partial E_p}-\frac{g_m{}^2 z_h\bar{z}_k}{(r_p+z_h)(r_p+\bar{z}_k)}\,\frac{\partial r_p}{\partial E_p}\right\}.$$

Instead of

$$_{h-k}a_2=\frac{1}{2}\,\frac{r_p}{r_p+{}_{h-k}z}\left\{\frac{\partial g_m}{\partial E_g}-\frac{2\mu_k\bar{z}_1}{r_p+{}_k\bar{z}_1}\,\frac{\partial g_m}{\partial E_p}+\frac{g_m{}^2{}_h z_1{}_k\bar{z}_1}{(r_p+{}_h z_1)(r_p+{}_k\bar{z}_1)}\,\frac{\partial r_p}{\partial E_p}\right\}.$$

This, however, does not affect the results or conclusions as (29) and following equations are not changed.

[3] E. T. Cunningham, Inc., New York City.

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BOOK REVIEW

A Bibliography of Electrical Literature. Serial No. 62; Publications from the Massachusetts Institute of Technology. 62 pages, paper bound; price 50c.

Contains publication data and outlines scope of the material included in publications and bulletins devoted to electrical information. It is useful as a guide for either the systematic or desultory study of published electrical engineering contributions from both American and foreign sources.

RALPH R. BATCHER †

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 1 *January, 1929*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a monthly list of references prepared by the Radio Laboratory of the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Bureau of Standards. The various periodicals can be consulted at large public libraries.

R100. Radio Principles

R007.1 Hooper, S. C. Considerations affecting the licensing of high frequency stations. Proc. I.R.E., **16**, 1240-51; September, 1928.
(Discussion based on study while with the Federal Radio Commission.)

R113 Edwards, S. W., and Brown, J. E. The use of radio field intensities as a means of rating the outputs of radio transmitters. Proc. I.R.E., **16**, 1173–93; September, 1928.
(A method described by which outputs of radio transmitting sets can be regulated by federal authority in terms of measured radio field intensities instead of watts in the transmitting set or antenna circuit. Work done on broadcast range only. Field intensity contour maps given for several stations.)

R113 Jansky, C. M. Jr. Some studies of radio broadcast coverage in the middle West. Proc. I.R.E., **16**, 1356–67; October, 1928.
(Statistical study based on reports to station WBBM.)

R113.2 Austin, L. W. Long-wave radio receiving measurements at the Bureau of Standards in 1927. Proc. I.R.E., **16**, 1252–57; September, 1928.
(Curves and tables give daylight signal intensities of a number of stations and strength of atmospheric disturbances.)

R113.4 Breit, G.; Tuve, M. A.; Dahl, O. Effective heights of the Kennelly-Heaviside layer in December, 1927 and January, 1928. Proc. I.R.E., **16**, 1236–39; September, 1928.
(Report on results of effective heights of Kennelly-Heaviside layer obtained Dec. 19, 1927 to Jan. 16, 1928 by the reflection method.)

R113.6 Hoag, J. B. and Andrew, V. J. A study of short-time multiple signals. Proc. I.R.E., **16**, 1368–74; October, 1928.
(An investigation of signals which may have travelled one or more times around the earth. Shows reflections of signals from region other than Heaviside layer.)

* Original manuscript received by the Institute, November 14, 1928.

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R125.6 Turlyghin, S. J. and Ponomareff, M. J. Zusammengesetze Rahmenantennen. (Combination of coil antennas). *Zeitschrift für Physik,* **9**, 356–64; 1928.

(Theoretical investigation of coil antennas and coil combinations, and coil antennas with reflectors.)

R125.6 Walmsley, T. Polar diagrams due to plane aerial reflector systems. *Experimental Wireless* (London), **5**, 575–77; October, 1928.

(Derivation of formulas by means of which the polar diagram in any vertical plane can be plotted. It is shown that the field behind the reflector is of considerable strength.)

R130 Crossley, A. and Page, R. M. A new method for determining efficiency of vacuum tube circuits. Proc. I.R.E., **16**, 1375–83; October, 1928.

(Method described for determining efficiency of vacuum-tube circuits for high and low frequencies by application of surface pyrometer indicating temperature of glass walls of tube.)

R131 Reed, M. The application of alignment charts to valve characteristics. *Experimental Wireless* (London) **5**, 571–74; October, 1928.

(The theory of alignment charts is given and applied to the tube equation for the anode and grid voltages and the plate current.)

R132 von Ardenne, M. The final or power stage of amplifiers. *Experimental Wireless* (London), **5**, 556–64; Oct., 1928.

(A discussion of the performance of the amplifier stage energizing a loudspeaker. The circuit is treated by means of the dynamic characteristic.)

R132 Medlam, W. B. Effect of anode-grid capacity in detector and low frequency amplifiers. *Experimental Wireless* (London), **5**, 545–55; October, 1928.

(An analytical treatment of the effect of the internal tube capacity for detector and low frequency amplifier circuits.)

R132 Watanabe, Y. Über den rückgekoppelten Verstärker. (On feedback amplifiers). *Zeitschrift für Hochfrequenztechnik,* **32**, 77–83; September, 1928.

(Equations for the amplification of amplifiers with electrostatic and magnetic feedback.)

R133 Grechowa, M. Zur Frage der Erzeugung kurzer elektromagnetischer Wellen. (On the production of short electro-magnetic waves). *Physikalische Zeitschrift,* **29**, 726–29; October 15, 1928.

(Based on an equation by Abraham; the case of the Barkhausen and the Gill and Morrell oscillations is studied with respect to their frequency.)

R133 Pfitzer, W. Die Selbsterregungsbedingungen bei Rückkopplungs- Röhrensendern für sehr kurze Wellen. (The condition for self excitation for short waves of tube generators with feedback) *Electrische-Nachrichten Technik,* **5**, 348–69; September, 1928.

(The oscillations for very short waves are investigated by taking into account the ground capacity and the interelectrode effects. It is shown that the circuit acts like a voltage divider. The feedback conditions can than be expressed by simple equations which give expressions for the natural oscillations.)

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R134 Terman, F. E. Some principles of grid-leak grid-condenser detection. Proc. I.R.E., **16,** 1384–97; October, 1928.

(Bases discussion on equivalent circuits, measures detection factors.)

R145.3 Wheeler, H. A. Simple inductance formulas for radio coils. Proc. I.R.E., **16,** 1398–1400; October, 1928.

(Two simple formulas presented which are patterned after an empirical formula derived by Prof. L. A. Hazeltine for inductance of simple types of radio-frequency coils.)

R200. Radio Measurements and Standardization

R201.2 Eisner, F. Über punktformige Aufnahme von Wechselstromkurven, insbesondere bei hoher Frequenz. (On the point by point method for tracing a.c. curves particularly at high frequency). *Archiv. für Elektrotechnik,* **20,** 473–502; Sept. 17, 1928.

(A three-electrode tube method is described for tracing wave shapes by the point to point method. The method seems good up to about 10 kc although results have been obtained up to 150 kc.)

R210 Westman, H. P. Frequency stability by magnetostriction oscillators. *QST,* **12,** 21–26; November, 1928.

(Comparison of oscillators and calibration of frequency meters by magnetostriction oscillators.)

R220 Graham, V. M. A gang capacitor testing device. Proc. I.R.E., **16,** 1401–03; October, 1928.

(Description of testing of gang condensers.)

R250 Coursey, P. R. On the application of condensers to the measurement of large radio frequency currents. *Experimental Wireless* (London), **5,** 565–71; October, 1928.

(The methods for measuring large radio-frequency currents are discussed and the method with two condensers in parallel is described in detail.)

R251.2 Colebrook, F. M. and Wilmotte, R. M. The design of non-contact thermo-junction ammeters. *Experimental Wireless* (London), **5,** 538–44; October, 1928.

(Contact and non-contact types of thermo-junction a.c. ammeters are compared. General principles are given of the designing of the non-contact type. Description of two designs of non-contact types embodying these principles with details as to performance. An account of a particular application as a known source of radio frequency voltage is given.)

R300. Radio Apparatus and Equipment

R326 Szekely, A. Über die einem Empfanger durch Erdung zugefuhrte Energie II. (On a receiving set, energy for which comes from the gound—II.) *Zeits. für Hochfrequenztechnik,* **32,** 83—86; September, 1928.

(The author finds that a coil antenna which has in addition a vertical lead-in to ground gives a larger current reading. The increase is due to the induction of the field on the vertical lead and is not due to the ground.)

R342 Rukop, H. Die elektrischen Eigenschaften der Rundfunksender-Vorverstarker im Hinblick auf ihre akustischen Qualities. (On the electrical properties of broadcast transmitter amplifiers with

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regard to their acoustic qualities). *Zeits. für Hochfrequenztechnik*, **32**, 86—93; September, 1928.

(Final paper of articles which appeared in this periodical 32, 18, 1928 and 32, 65, 1928. Deals with the frequency-amplitude characteristics of the amplifier.)

R342.15 Diamond, H. and Stowell, E. Z. Note on radio-frequency transformer theory. Proc. I.R.E., **16**, 1194–1202; September, 1928.

(General equations worked out which include the effect of the distributed capacity coupling existing between transformer windings.)

R382 Rucklin, R. Ein experimenteller Beitrag zum Spulenproblem. (An experimental contribution on the coil problem). *Archiv für Elektrotechnik*, **20**, 507–532; September 17, 1928.

(Experimental and theoretical investigation of single layer and pancake coils with respect to the critical frequency and current and voltage distribution. The theory includes also the contributions due to Rudenberg, Wagner, Boehm, von Steidingen, Rogowski, and Dreyfuss.)

R385.3 Schulgin, W. M. Der Wehnelt-Unterbrecher als Generator elektromagnetischer Schwingungen. (The Wehnelt interruptor as generator of electromagnetic oscillations.) *Physikalische Zeitschrift*, **29**, 724–26; October 15, 1928.

(Across a Wehnelt interruptor (platinum-lead) a resonator is connected and a voltage from a direct-current source impressed across the interruptor (platinum forms the negative pole). High-frequency oscillations are produced in the resonator circuit as in the arc generator.)

R385.5 Hartmann, C. A. Neuere Untersuchungen an Kohlmikrophonen. (New investigations on carbon microphones.) *Elektrische-Nachrichten Technik*, **5**, 344–47; September, 1928.

(Methods for testing microphones.)

R388 Sommerfeld, E. Über ein Kathodenoszillographen höher Spannungs-empfindlichkeit. (On a cathode-ray oscillograph of high voltage sensitivity.) *Archiv für Elektrotechnik*, **20**, 607–618; September 17, 1928.

(A cathode-ray oscillograph with accelerated grid and inside photographic arrangement with a sensitivity of 0.1 cm per volt is described.)

R388 Rogowski, W., Sommerfeld, E., and Wolman, W. Empfindlicher Glühcathodenoszillograph für Innenaufnahmen in einem Vorvakuum. (Sensitive cathode-ray oscillograph for photos inside the tube under a vacuum.) *Archiv für Elektrotechnik*, **20**, 619–24; September 17, 1928.

(The effect of the acceleration grid is brought out by several oscillograms. It is shown that slow electrons only slightly affect the photographic plate which is inside the tube and an increase of the speed of the electrons by fast transients does not help much. The slow moving electrons are therefore accelerated by a grid before falling in the plate. If an accelerating grid is not used, it is better to photograph the fluorescence with a camera outside the tube.)

R500. Applications of Radio

R521 Pratt, H. and Diamond, H. Receiving sets for aircraft beacon and telephony. Bureau of Standards Journal of Research, October, 1928. Research Paper No. 19. Reprint copies obtainable for 15 cents from the Supt. of Documents, Government Printing Office, Washington, D. C.

(Design details for three receiving sets of slightly different type developed for aircraft beacon and telephony. Brief discussion of the results of practical flight tests.)

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R526.1 Dunmore, F. W. Design of tuned reed course indicators for aircraft radiobeacon. Bureau of Standards Journal of Research, November, 1928. Research Paper 28. Reprint copies obtainable for 5 cents from the Supt. of Documents, Government Printing Office, Washington, D. C.

(Description of tuned reed indicators for aircraft radiobeacon, visual system.)

R526.1 Shangraw, C. C. Radiobeacons for transpacific flights. Proc I.R.E., **16**, 1203–35; September, 1928.

(Description of equi-signal radiobeacon system developed by the Air Corps and Signal Corps Aircraft Radio Labs. Application of these radiobeacons in the recent trans-pacific flights.)

R536 Jakosky, J. J. Electrical prospecting. Proc. I.R.E., **16**, 1305–55; October, 1928.

(Electrical geophysical methods for locating minerals.)

R582 Picture telegraphy. Post Office Engrs. *Jnl.* (London), **21**, 191–99; October, 1928.

(Résumé of various systems of picture telegraphy.)

R600. Radio Stations; Equipment, Operation and Management

R610 Pession, G. and Montefinale, G. Radiotelegraphic center at Rome (San Paolo). Proc. I.R.E., **16**, 1404–21; October, 1928.

(Description of station San Paolo including new short-wave transmitting set.)

R800. Non-Radio Subjects

534 Trendelenburg, F. Zusammenfassenden Bericht: Über neuere akustische und insbesondere elektroakustische Arbeiten. (Summary report: On new acoustic and especially electrical acoustic papers.) *Zeits. für Hochfrequenztechnik*, **32**, 94–99; September 1928.

(In this continuation of a series of papers the acoustic field in space is discussed. The treatment considers reverberation such as originally given by W. C. Sabine and lately by F. R. Watson.)

534 Die Tonerzeugung durch Spitzen an hohen Wechselpotential und ihre Verwendung als membranloser Lautsprecher. (The sound production by means of needle gaps and high alternating potential and its application to a loudspeaker without a diaphragm.) *Naturwissenschaft*, **16**, 795–96; October 19, 1928.

(A method is described where by means of a needle gap and a high voltage across it, sounds are produced. This principle is then applied to a new type of a loudspeaker.

534.3 Harrington, E. A. The vibration of tuning forks. *Jnl.* Optical Soc. and Rev. of Scientific Instruments, **17**, 224–39; September, 1928.

(Theoretical and experimental study of electrically driven tuning forks and bars. Results: The deflection of the prongs approximately equal to equivalent length and only about 3.5 per cent of the total energy converted into sound.)

534.83 Service, J. A. Radio acoustic position finding in hydrography. *Journal* A.I.E.E., **47**, pp. 670–74; September, 1928.

(In this method the sound of a bomb alongside the ship whose position is to be determined affects a microphone at the shore and by means of it sends out a radio signal

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which is recorded on the same chronograph as the report of the bomb on the ship. The time interval between sound report and radio signal gives a means for calculating the distance to the shore. In addition, a device for quickly plotting the field sheet is described.)

537.65 Skellet, A. M. Visual method for studying modes of vibration of quartz plates. *Jnl.* Optical Soc. of America and Rev. of Scientific Instruments, **17**, 308–17; October, 1928.

(Experimental study of the different modes of resonator frequencies of quartz plates. The glow discharge method of Giebe and Scheibe is used.)

538 Masiyama, Y. On the magnetostriction of a single crystal of nickel. Science Reports of Tohoku University, Japan, **17**, 945–61; August, 1928.

(Longitudinal and transverse effects due to magnetostriction were studied in a crystal of nickel whose shape was that of an oblate ellipsoid. When the ellipsoid was in different positions of a magnetic field, the longitudinal effect produces a contraction. The transverse effect is the reverse.)

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VOLUME 17 FEBRUARY, 1929 NUMBER 2

PROCEEDINGS
of
The Institute of Radio Engineers

1929 CONVENTION
Washington, D. C.
May 13-15

Published Monthly By
THE INSTITUTE OF RADIO ENGINEERS
Publication Office: 450-454 Ahnaip St., Menasha, Wis.

BUSINESS, EDITORIAL AND ADVERTISING OFFICES
33 West 39th Street, New York, N.Y.

Subscription $10.00 per Annum in the United States
$11.00 in all other Countries

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Institute of Radio Engineers

Forthcoming Meetings

FOURTH ANNUAL CONVENTION
Washington, D. C., May 13-15, 1929

ATLANTA SECTION
Atlanta, Georgia, February 8, 1929

CLEVELAND SECTION
Cleveland, Ohio, February 15, 1929

NEW YORK MEETING
New York, N. Y., February 6, 1929 and March 6, 1929

PHILADELPHIA SECTION
Philadelphia, Penna., February 19, 1929

PITTSBURGH SECTION
Pittsburgh, Penna., February 21, 1929

TORONTO SECTION
Toronto, Canada, February 20, 1929

WASHINGTON SECTION
Washington, D. C., February 14, 1929

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 17 February, 1929 Number 2

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The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

Subscription rates to the PROCEEDINGS for the current year are received from non-members at the rate of $1.00 per copy or $10.00 per year. To foreign countries the rates are $1.10 per copy or $11.00 per year.

Back issues are available in unbound form for the years 1918, 1920, 1921, 1922, 1923, and 1926 at $9.00 per volume (six issues) or $1.50 per single issue. Single copies for the year 1928 are available at $1.00 per issue. For the years 1913, 1914, 1915, 1916, 1917, 1918, 1924, and 1925 miscellaneous copies (incomplete unbound volumes) can be purchased for $1.50 each; for 1927 at $1.00 each. The Secretary of the Institute should be addressed for a list of these.

Discount of twenty-five per cent on all unbound volumes or copies is allowed to members of the Institute, libraries, booksellers, and subscription agencies.

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Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership, etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

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Changes of address to affect a particular issue must be received at the Institute office not ater than the 15th of the month preceding date of issue. That is, a change in mailing address to be effective with the October issue of the PROCEEDINGS must be received by not later than September 15th. Members of the Institute are requested to advise the Secretary of any change in their business connection or title irrespective of change in their mailing address, for the purpose of keeping the Year Book membership catalog up to date.

The right to reprint limited portions or abstracts of the papers, discussions, or editorial notes in the PROCEEDINGS is granted on the express condition that specific reference shall be made to the source of such material. Diagrams and photographs published in the PROCEEDINGS may not be reproduced without making special arrangements with the Institute through the Secretary.

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Entered as second class matter at the Post Office at Menasha, Wisconsin.

Acceptance for mailing at special rate of postage provided for in the Act of February 28, 1925, embodied in paragraph 4, Section 412, P. L. and R. Authorized October 26, 1927.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered, and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer de visions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportions.

Abbreviations—Write a.c. and d.c., kc, μf, $\mu\mu$f, emf, mh, μh, henries, abscissas, antennas Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right, in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

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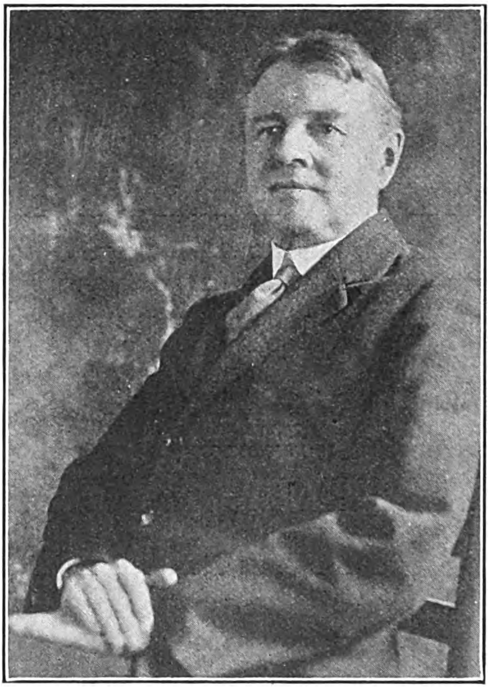

L. W. Austin

President of the Institute of Radio Engineers, 1914

L. W. Austin was born at Orwell, Vermont, October 30, 1867. He received the degree of Bachelor of Arts from Middlebury College in 1889, and the Ph.D. degree from the University of Strassburg in 1893. From 1893 to 1901 he was instructor and assistant professor at the University of Wisconsin, following which he did research work at the University of Berlin in 1901–1902. Since 1904 Dr. Austin has been with the Bureau of Standards, Washington, D. C.; head of U. S. Naval Radio Research Laboratory, 1908–1923; chief of Radio Physics Laboratory, 1923 to date.

Dr. Austin was President of the Institute in 1914, served on the Board of Direction from 1915 to 1917, and was awarded the Institute Medal of Honor in 1927. His contributions to the Proceedings have been frequent. Dr. Austin was elected to Associate membership in the Institute in 1913, transferred to Member grade in 1913 and to the Fellow grade in 1915.

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INSTITUTE NEWS AND RADIO NOTES

Fourth Annual Convention of the Institute

Washington, D. C., May 13th to 15th

The tentative program for the 4th Annual Institute Convention, to be held in Washington, D. C. on May 13 to 15th has been arranged. The Convention Committee Chairmen are as follows: C. B. Jolliffe, Convention Chairman; T. Mc. L. Davis, Registration and Arrangement; S. S. Kirby, Trips; F. P. Guthrie, Dinner and Entertainment; Mrs. L. W. Austin, Ladies; A. E. Kennelly, Fellowship; W. G. H. Finch, Publicity.

Convention Headquarters will be in the Mayflower Hotel where a number of rooms are being set aside for reservation by Institute members. Registration, Committee meetings, several functions for the ladies, the banquet, etc., will be held in the Mayflower, which is conveniently located with respect to the U. S. Chamber of Commerce Building in which the technical sessions are to be held, and can be reached readily either by auto, streetcar or bus.

The tentative program is as follows:

SUNDAY, MAY 12

2 P.M. to 6 P.M.—Registration at the Mayflower Hotel.

MONDAY, MAY 13

8:00 A.M.	Registration at the Mayflower Hotel.
10:00 A.M.	Opening Session; Speeches by President A. Hoyt Taylor, F. P. Guthrie and C. B. Jolliffe; Symposium on "Technical Problems of Radio Regulation."
11:00 A.M.	Meeting of representatives of Sections.
12:30 P.M.	Lunch.
1:15 P.M.	Inspection Trip leaving for Naval Research Laboratory, Bellevue, Anacostia, D.C., returning to the Mayflower Hotel at 5:15 P.M.
8:00 P.M.	Popular Lecture by Professor M. I. Pupin.

TUESDAY, MAY 14

9:00 A.M.	Symposium on Photo Radio.
10:30 .M.A	Inspection trip to Arlington Radio Station via Potomac Park, Arlington, and Arlington Cemetery.
12:30 P.M.	Lunch.

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1:15 P.M. Inspection trip to Bureau of Standards via 16th Street and Rock Creek Park.

4:15 P.M. Short speech by Director of Bureau of Standards.

4:30 P.M. Tea at the Bureau of Standards, returning to Headquarters at 5:30 P.M.

7:30 P.M. Banquet at the Mayflower Hotel, including speeches by President A. Hoyt Taylor, several Institute members, presentation of Institute prizes, talking moving picture feature, and facilities for dancing.

WEDNESDAY, MAY 15

Annual meeting of the American Section, International Scientific Radio Union, to which members of the Institute are invited. This meeting will consist of some twenty technical papers, averaging fifteen minutes each. The papers will cover developments in the fields of work of the Union's technical Committees. Besides the technical papers on particular developments, there will be a general report on the status of the field by the chairman of each technical committee, as follows: radio measurements, J. H. Dellinger; measurements of interference, E. F. W. Alexanderson; wave propagation, L. W. Austin; wave direction, G. Breit; phenomena above 3000 kilocycles, A. H. Taylor; atmospherics, H. T. Friis; cooperation, A. E. Kennelly; radio physics, E. L. Chaffee.

The program of specific papers in these fields will be published in a later issue. It is expected that preprints of the papers will be available in advance of the meeting.

Change in Proceedings Format

With the January issue the arrangement of PROCEEDINGS pages was changed with a view to making the various departments of each issue more readily available to members. With the April issue it is contemplated that the *type* size of each page will be increased to allow approximately fifteen per cent additional reading matter to be printed on each page.

Comments from the membership, both as to the typographical arrangement of the PROCEEDINGS and as to its contents, will be appreciated by the Board of Direction. Constructive criticism is always welcomed in the endeavor to make the PROCEEDINGS as attractive as possible to all grades of membership.

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January Meeting of the Board of Direction

The following were present at the meeting of the Board of Direction of the Institute held on January 2nd in the Institute office: Alfred N. Goldsmith, President; L. E. Whittemore, Vice-President; Melville Eastham, Treasurer; Ralph Bown, Junior Past President; Arthur Batcheller, W. G. Cady, J. H. Dellinger, R. A. Heising, J. V. L. Hogan, R. H. Marriott, and John M. Clayton, Secretary.

The count of the ballots for 1929 Officers and new Board members brought the following results: President, A. Hoyt Taylor; Vice-President, Alexander Meissner; Managers elected for three-year terms, Arthur Batcheller and C. M. Jansky, Jr.

The Board appointed the three Managers with one-year terms as follows: Lewis M. Hull, R. H. Marriott, and L. E. Whittemore.

The Board approved the action of the Committee on Admissions through the election or transfer of the following: elected to the Fellow grade: Lt. Cmdr. T. A. M. Craven; transferred to the Member grade: Avery G. Richardson, Harvey Meisenheimer, and L. W. Branch; elected to the Member grade: Knox C. Black, W. B. Morehouse, and Albert Kofes.

Seventy-eight Associate members and eight Junior members were elected.

Paul A. Greene, of the Columbia Broadcasting System, was appointed to membership on the Committee on Broadcasting.

Notices of New York Meetings

As authorized by the Board of Direction, and effective immediately, notices of regular New York meetings of the Institute will, in the future, be sent only to members residing within the states of Connecticut, Delaware, Maryland, Massachusetts, New Jersey, New York, Pennsylvania, and District of Columbia, unless specific request be made once a year by individual members in other states.

1928 Bound Volumes

The 1928 volume of the PROCEEDINGS is available in a handsome blue buchram binding at $9.50 (one dollar additional for foreign postage). Only a limited number of volumes have been

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bound. Members desiring to obtain the 1928 volume in bound form are also cautioned that unbound copies for the entire year are *not* available.

The above prices are to members of the Institute.

Standard Frequency Transmissions by the Bureau of Standards

The Bureau of Standards announces a new schedule of radio signals of standard frequencies, for use by the public in calibrating frequency standards and transmitting and receiving apparatus. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington which went into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to 1,000 miles from the transmitting station.

The transmissions are by continuous-wave radiotelegraphy. The modulation which was previously on these signals has been eliminated. A complete frequency transmission includes a "general call" and " standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues for about 2 minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about 4 minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

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Radio Transmissions of Standard Frequency; Schedule of Frequencies in Kilocycles

Eastern Standard Time	March 20	April 22	May 20	June 20	July 22
10:00–10:08 P.M.	1500	4000	125	550	1500
10:12–10:20	1700	4500	150	600	1700
10:24–10:32	2250	5000	200	700	2000
10:36–10:44	2750	5500	250	800	2300
10:48–10:56	2850	6000	300	1000	2700
11:00–11:08	3200	6500	375	1200	3100
11:12–11:20	3500	7000	450	1400	3500
11:24–11:32	4000	7300	550	1500	4000

Report of Federal Radio Commission

The annual report of the Federal Radio Commission to the Congress of the United States for the year ending June 30, 1928, together with a supplementary report for the period from July 1, 1928 to September 30, 1928, can be procured from the Superintendent of Documents, Government Printing Office, Washington, D. C., for twenty-five cents per copy.

The report is a work of some two hundred and sixty pages and is divided into three parts; part 1 contains a statement of the personnel and organization of the Commission; part 2 is devoted to the Commission's work with regard to the broadcast band; part 3 is devoted to low and high-frequency bands. The report also contains copies of all General Orders issued by the Commission between July 1, 1927 and October 26, 1928.

Institute Meetings

Atlanta Section

The Atlanta section held a meeting in the Chamber of Commerce Building, Atlanta, Ga., on January 9th. Major Van Nostrand, chairman of the section, presided.

A paper, "Modern Merchandising methods in the Radio Field," was presented by Pierre Boucheron, of the Radio Corporation of America.

Seventeen members of the section attended the meeting.

The February 8th meeting will be held in the Chamber of Commerce Building. At this meeting J. K. Clapp, of the General Radio Company, will present a paper "A Convenient Method for Referring Secondary Standards to a Standard Time Interval."

Boston Section

A meeting of the Boston section was held in the Cruft Laboratory, Harvard University, Cambridge, Mass., on December 12th. This was a joint meeting with the Boston Signal

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Post. Dr. G. W. Pierce and D. S. Boyden presided. Major Richard H. Ranger, of the Radio Corporation of America, presented a paper, "Recent Developments in Photoradio." Messrs. Sewall Cabot, A. F. Murray, P. S. Bauer, H. W. Lamson, and R. W. Hall participated in the discussion of the paper.

One hundred and sixty-two persons were present at the meeting.

Buffalo-Niagara Section

A meeting of the Buffalo-Niagara section was held on November 14th in Edmund Hayes Hall, University of Buffalo. L. C. F. Horle, chairman of the section, presided.

John F. Morrison, of Radio Station WKBK, presented a paper, "The Modulation System of WKBK."

Edward Stanko, of Station WGR, presented a paper, "Broadcast Antenna Characteristics," and Francis D. Bowman presented a paper, "Present Day Broadcasting from the Advertiser's Viewpoint."

Forty members of the section attended the meeting.

At the December 12th meeting of the Buffalo-Niagara section D. E. Replogle presented a paper, "Television." Messrs. Horle, Hector, Johnson, Smith, Freck, and Henderson participated in the discussion which followed.

Sixty members of the section attended the meeting.

On January 17th L. Grant Hector, of the University of Buffalo, will present a paper, "Apparent Equality of Loudspeaker Output at Various Frequencies."

Chicago Section

A meeting of the Chicago section was held January 11th at the Electric Club, Chicago. John H. Miller, chairman of the section, presided.

Fred H. Schnell, of the Burgess Battery Company, presented a paper, "Short Wave Aircraft Radio." The paper explained the work of the Burgess Battery Company on dry cell operated transmitters and receivers for aircraft. A rather detailed analysis was given of the requirements for equipment for radio service on airplanes, and it was shown that this varied to a wide extent. Actual tests were given of light weight transmitters. A description of the apparatus installed on the Rockford Plane, which flew to Greenland, was given. Lantern slides covering this paper were shown.

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Following the presentation of the paper Messrs. Oxner, Miller, Wilcox, Minnium, and Kennedy participated in the discussion.

The new officers of the Chicago section were elected as follows: H. E. Kranz, chairman; B. J. Minnium, vice-chairman; John H. Miller, secretary-treasurer.

Cleveland Section

A meeting of the Cleveland section was held in the Case School of Applied Science in Cleveland, Ohio, on December 14th. Professor John R. Martin, chairman of the section, presided and presented a paper on "The Use of the Phonodeik in the Study of Loud Speakers and the Electric Pick-up." The talk was illustrated with lantern slides and was accompanied by a demonstration of a special model of the phonodeik. Professor Martin emphasized the difficulty in making tests by ear, as the personal factor of likes and dislikes affects the observer. The advantage of a sound photographing device in comparing tones throughout the audible range lies in its ability to actually record the air motion or its electrical counterpart at any point in its transmission. The direct method eliminates the personal element in studying sound reproducing, amplifying, and pick-up devices.

Extensive discussion followed the presentation of the paper. The report of the Nominating Committee for 1929 officers of the Cleveland section was unanimously adopted. These officers are: B. W. David, chairman; Ralph Farnham, vice-chairman; D. Schregardus, secretary-treasurer; D. M. Ward, chairman of Committee on Membership; Prof. J. R. Martin, chairman of Program Committee.

Sixty-eight members of the section attended the meeting.

Connecticut Valley Section

A meeting of the Connecticut Valley section was held on January 18th. K. S. Van Dyke, vice-chairman of the section, presided. J. K. Clapp, of the General Radio Company, presented a paper on "Short Wave Short Distance Radio Transmission."

The paper dealt with the electronization of the atmosphere and ideal paths of transmission. Data of a series of tests run between Boston and South Dartmouth, Mass., working on various wavelengths around 40 meters which gives a very logical

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explanation of the so-called "skip distance," "flutter zone," and other short-wave phenomena were presented. Results of tests made with an antenna which could be rotated or placed at different angles with the earth were shown.

Messrs. Laport, Van Dyke, and others participated in the discussion which followed.

The officers for 1929 were elected and are as follows: Q. A. Brackett, chairman; E. A. Laport, vice-chairman; F. C. Beekley, secretary-treasurer.

Thirty-seven members of the section were present at this meeting.

Detroit Section

A meeting of the Detroit section was held on November 23rd in the Detroit News Building.

Earle D. Glatzel, chairman of the section, presided. Professor Roy S. Glasgow, of Washington University, St. Louis, Mo., presented a paper on "Some Recent Developments in Broadcast Receiving Set Design."

The paper discussed factors affecting selectivity, sensitivity, and fidelity of receiving sets, after which a new type of receiver employing a band pass filter tuning system followed by a fixed transformer coupled radio-frequency amplifier was described. Design of the filter to give a constant width of received band throughout the broadcast band range was discussed. The radio-frequency amplifier utilizes variation of the input capacitance of a vacuum tube with the plate circuit reactance automatically tuning the amplifier circuits to the frequency being received.

Forty-five members of the Institute and guests were present at this meeting.

The Detroit section held its December meeting on the 21st of the month in the Detroit News Building, 615 Lafayette Blvd. E. D. Glatzel, chairman of the section, presided.

C. G. Hall, of the Cardon Corporation of Jackson, Michigan, spoke on "Manufacturing Production of Vacuum Tubes." A detailed description of the processes involved in turning out vacuum tubes at the rate of 10,000 a day was given, together with an explanation of methods and apparatus used in manufacturing the glass parts of the tube, making and assembling the elements, degassing, sealing, and evacuating, and the final testing of the product.

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Los Angeles Section

On November 19th a meeting of the Los Angeles section was held in the Elite Cafe, Los Angeles. Don C. Wallace, chairman of the section, presided. Lt. Commander Lowell Cooper, U.S.N., presented a talk on "Naval Radio Procedure and Practice." The address covered the general training, routine, and practice employed in naval radio control as well as the method of transmitting messages from different squadron groups to the flag ship. A superficial description of the types of radio equipment used was also given.

Lieutenant Dean Farrand, U. S. Army reserve, presented a talk on "Aircraft Beacons and Radio Equipment" in which were described the radio beacons in use along the coast and in Hawaii. Blackboard diagrams of their field distribution were given, and the speaker described some of the difficulties and problems of commercial aviation and the limitations imposed on radio equipment. The necessity for visual rather than oral indicators for the pilot was stressed.

W. S. Halstead, U. S. Forestry Service, gave a talk on "Radio as Applied to Fire Control and Forestry." By means of lantern slides a number of interesting views by a combined radio transmitter and receiver suitable for portable service in forestry and fire control work were shown. The talk included various technical data descriptive of the radio equipment and operating methods.

Forty-six members of the section attended the meeting.

The following officers were elected for the coming season: Thomas F. McDonough, chairman; Theodore C. Bowles, vice-chairman; W. W. Lindsay, secretary-treasurer.

The section's Board of Directors was elected as follows: Don C. Wallace, Theodore C. Bowles, Captain F. E. Pierce, Robert B. Parrish, Lee Yount, W. W. Lindsay, and Thomas F. McDonough.

New Orleans Section

A meeting of the New Orleans section was held on December 29th, presided over by Pendleton E. Lehde, chairman.

Anton A. Schiele presented a paper, "Control of Street Lights by Means of Tuned Circuits." Messrs. Jones, Gallo, and Mackie participated in the discussion which followed. Twelve members of the New Orleans section attended the meeting.

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The New Orleans section held a meeting January 10th at which Chairman Pendleton E. Lehde presided.

Professor C. M. Jansky, Jr. read a paper entitled "Location of Salt Domes by Geo-Physical Explorations." The following members took part in the discussion that followed: Professor Ricker, J. N. DuTreil, and George Deiler.

Forty members and guests attended the meeting.

On February 11th a meeting of the New Orleans section will be held at which time J. K. Clapp, of the General Radio Company, will present a paper on "A Convenient Method for Referring Secondary Standards to a Standard Time Interval.

New York Meeting

A New York meeting of the Institute was held on January 2nd in the Engineering Societies Building, 33 West 39th Street, presided over by Alfred N. Goldsmith. V. Zworykin, of the Westinghouse Electric and Manufacturing Company, presented a paper on "Facsimile Picture Transmission." This paper will be published in an early forthcoming issue of the PROCEEDINGS.

Following its presentation the following participated in the discussion: Messrs. Goldsmith, McCullough, Gallagher, Bonn, Nyman, Ballantine, Hagglung, Dewhirst, Shannon, and Ranger.

Three hundred members of the Institute and guests attended this meeting.

Pittsburgh Section

The January 15th meeting of the Pittsburgh section was held in the Fort Pitt Hotel. L. A. Terven, vice-chairman of the section, presided. J. G. Allen and A. Mag presented "A Symposium on Radio Interference."

The paper showed how the source of noise is traced from the high voltage down to the original house current. It was stated that less than fifteen per cent of the radio interference cases reported were due to public utility companies. Electrical appliances are responsible to a great degree for the interference caused. It was also stated that appliances should be made "interference proof" at the factory. A demonstration of "noise makers" was given including universal motors, heating pads, tree grounds, etc. Messrs. Froelich, McKinley, Horn, and Terven participated in the discussion which followed.

Twenty members of the section attended the meeting.

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On February 21st there will be a meeting of the Pittsburgh section in the Fort Pitt Hotel. J. K. Clapp, of the General Radio Company, will present a paper on "A Convenient Method for Referring Secondary Standards to a Standard Time Interval."

The Pittsburgh section held a joint meeting with the Pittsburgh section of the American Institute of Electrical Engineers in the Chamber of Commerce Building on December 11th.

H. A. Iams presented a paper on "Facsimile Picture Transmission."

Twenty-one members of the Institute were present.

Rochester Section

The Rochester section held a meeting on November 23rd in the Rochester Chamber of Commerce Building. A. B. Chamberlain, chairman of the section, presided.

Three minute speeches were made by J. P. Boylan on "Telephony"; F. W. Reynolds on "Broadcasting"; C. E. K. Mees on "Research"; A. T. Haugh on "Merchandising"; C. L. Cadle on "Power."

L. B. F. Raycroft, vice-president in charge of Radio Division of the National Electrical Manufacturers' Association, talked on "The Radio Industry."

One hundred and thirty-two members attended the meeting.

On December 7th a meeting of the Rochester section was held in the Sagamore Hotel, Rochester. A. B. Chamberlain, chairman of the section, presided.

The meeting was addressed by Conan A. Priest, of the General Electric Company, on the subject of "Short Waves." The speaker presented a summary of recent progress in the development and use of short waves explaining the various operating and economic advantages of their use. This was a joint meeting between the Rochester section of the American Institute of Electrical Engineers, the Rochester Engineering Society, and the Rochester section of the Institute of Radio Engineers.

Sixty-three members of these societies attended the meeting.

San Francisco Section

On November 21st a meeting of the San Francisco section was held in the Bellevue Hotel, 505 Geary Street, San Francisco. Leonard F. Fuller, chairman of the section, presided.

Major Henry L. Dunn, Signal Corps Reserve, presented a talk on "The Radio Intelligence Service in the World War."

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Major Dunn spoke of many phases of the Intelligence Service of the U. S. Army during the late war. The means of obtaining information was given, the source of information being from spies, enemy prisoners, airplane observers, and observers using electrical methods. The use to which the information was put in keeping track of enemy forces was outlined. Methods of securing information directly from German telephone lines by actual contact, induction, and grounded lines with low-frequency amplifiers was described in detail. The development of radio compass intercepts stations in locating secret transmitters was mentioned. The talk was illustrated with lantern slides and was made interesting by an account of many humorous incidents.

Twenty-nine members and guests attended the meeting.

Seattle Section

A meeting of the Seattle section was held on November 30th in the Washington Engineer's Club, Arctic Building, Seattle. Walter A. Kleist, chairman of the section, presided.

Austin V. Eastman presented a paper on "Four-Electrode Tube as an Audio Amplifier." The speaker outlined some of the short-comings of three-electrode tubes as audio-frequency amplifiers and showed the need for the four-electrode type. He briefly outlined the use of the 222-type tube as a radio-frequency amplifier and next discussed this tube as an audio-frequency amplifier. Characteristic curves were drawn and a description of laboratory measurements of the overall voltage gain of a stage of impedance coupled amplification using the 222-tube as a space charge tube was given. Curves were presented showing the gain as the frequency was varied from less than a hundred cycles to six thousand cycles. The limit of input voltage with a consideration of distortion effect was explained.

Messrs. Tolmie, Kleist, and Willson discussed the paper.

Thirty-two members of the section attended this meeting.

Toronto Section

The November meeting of the Toronto section was held in the Electrical Building, University of Toronto, on the 23rd of the month, presided over by V. G. Smith.

R. H. Langley, of the Crosley Radio Corporation of Cincinnati, Ohio, presented a paper on "First National Broadcasting Station."

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The paper described in some detail the new 50 kw station placed in operation in October of 1928 by the Crosley Radio Corporation. Description of the transmitter and the antenna lay-out was given. Following the presentation of the paper Messrs. Lowry, Hackbusch, Smith, Pipe, and others participated in its discussion.

On January 16th the Toronto section held a meeting in the Electrical Building, University of Toronto. A. M. Patience, chairman, presided.

J. M. Thompson, of the Ferranti Electric Company, presented a paper on "Transient Conditions in Output Transformers." Following the presentation of the paper Messrs. Cline Hackbush, Bayley, and others participated in its discussion.

Forty-five members of the section were present at this meeting.

On February 20th there will be a meeting of the Toronto section in the Electrical Building, University of Toronto, at which time Mr. Thompson, of the Marconi Company, will present a paper on "Marconi Beam System."

A meeting of the Toronto section was held in the Electrical Building, University of Toronto, on December 19th, presided over by A. M. Patience.

Lieut. W. L. Laurie, Royal Canadian Signals, presented a talk on radio activities of Hudson Straits Expedition which was sponsored by the Canadian Government in July of 1927 and returned in the fall of 1928. The members of the expedition established three bases along the coast, using radio to keep them in touch with civilization at Ottawa. Three aeroplanes were used in the expedition and were equipped with lightweight radio sets built by the Air Force Laboratories in Ottawa. The sets were operated by dry batteries for filament supply and wind driven generators for high voltages. Both telephony and telegraphy were used with a maximum radius varying between 300 and 900 miles. Messrs. Pollack, Choat, Lowry, and others participated in the discussion which followed.

Forty-five menbers of the section were present at the meeting.

Washington Section

On December 13th a meeting of the Washington section was held in the Continental Hotel, North Capitol Street, N.W., Washington, D. C. F. P. Guthrie, chairman of the section, presided.

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E. O. Hulburt, of the Naval Research Laboratory, presented a paper, "Recent Developments in the Theory of High Atmospheres of the Earth," which included a brief summary of four papers which are to be published on the following subjects: "A Theory of Auroras and Magnetic Storms," "The Ion Ring Around the Earth at a Distance of about 45,000 Kilometers," and "Comets and Magnetic Storms."

Following the presentation of the paper A. H. Taylor, H. G. Dorsey, August Hund, W. H. Crew, G. D. Robinson, F. P. Guthrie, W. B. Burgess, and Charles Speaker participated in its discussion.

Preceding the meeting a dinner was held at which thirty-nine members and guests were present. The meeting was attended by seventy members of the section.

The Washington section held a meeting in the Continental Hotel, Washington, D. C. on January 10th. F. P. Guthrie chairman of the section, presided.

L. A. Hyland, of the Aircraft Section, Naval Research Laboratory, presented a paper "The Elimination of Ignition Interference to Aircraft Radio Reception."

The paper gave an outline of the procedure followed in the past few years to eliminate interference from aircraft radio reception. The various forms of interference were pointed out and their effects and remedies discussed. The several forms mentioned consisted of wind noises, propellor whine, exhaust noises and motor vibration, all of which produced microphonic disturbances from vibration of the plane or the surrounding air, and finally the ignition interference which was considered as the worst offender. The use of tuned audio systems were recommended for reducing the microphonic disturbances and shielded ignition systems for ignition interference elimination. Particular reference was made to recent shielded spark plug development, which has resulted in a design that not only makes shielding practical but embodies features that should increase the life of the plug with a possible increase in ignition efficiency.

Messrs. Dorsey, Price, Robinson, Guthrie, Burgess, Mirick, and Jackson participated in the discussion which followed.

Fifty-two members of the section attended this meeting.

The February 14th meeting will be held in the Continental Hotel and J. K. Clapp, of the General Radio Co., will present a paper "A Convenient Method for Referring Secondary Standards to a Standard Time Interval."

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Committee Work

Committee on Standardization

During the first few months of 1928, the Committee on Standardization, L. E. Whittemore, chairman, continued as during the previous year, with the formulation of material by the subcommittees which had been organized. These subcommittees and their chairmen were as follows: Subcommittee on Vacuum Tubes,—L. A. Hazeltine, chairman, 1927, C. B. Jolliffe, chairman, 1928; Subcommittee on Receiving Sets,—J. H. Dellinger, chairman, 1927, E. T. Dickey, chairman, 1928; Subcommittee on Electro-Acoustic Devices,—R. H. Manson, chairman; Subcommittee on Circuit Elements,—H. M. Turner, chairman; Subcommittee on Power Supply,—W. E. Holland, chairman; Subcommittee on Use of the Transmission Unit,—J. V. L. Hogan, chairman; Subcommittee on Bibliography,—C. A. Wright, chairman.

The material prepared by these subcommittees up to May 25, 1928 was printed in a Preliminary Draft Report bearing that date. Notices of the copies of this Preliminary Report available at that time were printed in the Proceedings. Copies were also distributed at the conventions of the two associations of radio manufacturers held in Chicago, in June, and were sent to others who were likely to be able to give constructive comments.

Comments and criticisms of this report were compiled and sent to members of the committee for consideration prior to a series of meetings held in New York during the latter part of 1928. Meetings were held on October 2, October 16, November 8, and December 6, 1928.

At these meetings consideration was given to the material contained in the Preliminary Report and to the comments and additional material submitted for adoption. A committee on form and arrangement, consisting of Haraden Pratt, E. T. Dickey, and the chairman, was authorized to put the material adopted by the committee into final form for submission to the Board of Direction of the Institute. The scope of the report is indicated by the following items included in the table of contents of the preliminary draft:

Definitions of Terms Used in Radio Engineering
Waves and Wave Propagation, Transmitting, Receiving, Vacuum Tubes, Circuit Elements, Properties, Antennas, Direction Finding, Electric-Acoustic Devices

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Standard Graphical Symbols
Letter Symbols; Vacuum-Tube Notation
Standard Methods of Measuring the Important Characteristics of Vacuum Tubes
Performance Indices of Electro-Acoustic Devices
Standard Methods of Measuring Properties of Circuit Elements; Chart and Bibliography

The Institute had continued to have the cooperation, through membership on the committee, of representatives of the American Institute of Electrical Engineers, the radio division of the National Electrical Manufacturers' Association, and the Radio Manufacturers' Association. The policy formulated in 1927 under which the field of radio standardization has been divided between the Institute of Radio Engineers on the one hand and the manufacturers' associations on the other has been continued. Partly as a result of this policy, it is believed that there is a growing support of the work of the Institute by other branches of the radio industry.

Through interlocking committee personnel and by other means every effort has been made to take advantage of the results of related work done by other organizations and to avoid duplication of effort.

In accordance with the plan adopted by the Board of Direction it is anticipated that the final report will be published as a part of the 1929 Year Book to be issued early in 1929.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED JANUARY 2, 1929

Elected to the Fellow grade

Dist. of Columbia	Washington, Naval Communications, Navy Dept	Craven, T. A. M.

Transferred to the Member grade

Michigan	Muskegon, 1390 Palmer Avenue	Richardson, Avery G.
New York	New York City, 195 Broadway Room 1607	Misenheimer, Harvey N.
Hawaii	Hilo, Box 923	Branch, L. W.

Elected to the Member grade

Massachusetts	Cambridge, 50 Kirkland Street	Black, K. Charlton
New Jersey	Bloomfield, 120 Berkeley Avenue	Morehouse, W. B.
Germany	Berlin-Charlottenburg, Cauerstrasse 19III	Kofes, Albert

Elected to the Associate grade

Arkansas	Camden, Box 661	Tagart, Sam W.
	Texarkana, 2119 Hickory Street	Wallace, L. E.
California	Los Angeles, 947 Francisco Street	Heimberger, Albert E.
	San Francisco, 378 Golden Gate Ave., Apt. 235	Kelsey, Karl D.
Connecticut	Hartford, 33 Eastview Street	Le Conche, Carl
Dist. of Columbia	Washington, 3131 Newton Street, N.E.	Carroll, Thomas D.
Illinois	Chicago, 4141 Fifth Avenue	Felio, Orin J.
	Chicago, 2530 Kedzie Blvd	Krueger, Alfred
	Chicago, 2530 Kedzie Blvd	Wilm, C. F.
	Galesburgh, 895 West Main Street	Mead, Leo R.
Indiana	North Manchester, 702 North Walnut St.	Grove, Claude C.
	Richmond, Box 6, Earlham College	Hickman, Roger W.
	Valparaiso, 825 Lincolnway	Swanson, Carl R.
Kansas	Topeka, c/o Radio Station WIBW	Herider, Ernest D.
	Wichita, 104 Severdale Apts	Mitchell, Ray H.
Kentucky	Covington, 3410 Church Street	Fraasa, C. F.
Louisiana	New Orleans, c/o Tropical Radio Tel. Co. 321 St. Charles Street	Allston, W. F.
	Port Barre	Futral, John E.
Maine	Bangor, 331 Center Street	Creamer, W. J., Jr.
Massachusetts	Boston, 200 Huntington Avenue	Browne, Monte C.
	Boston, 468 Massachusetts Ave	Wass, Howard H.
	Cambridge, 28 Gorham Street	Shen, Pao-guay
	Cambridge, 28 Gorham Street	Tsao, T. C.
	Rockport, 7 Gott Street	Mills, William P.
	Springfield, 38 Greenacre Square	Nystrom, Raymond A.
Michigan	Ann Arbor, 520 Forest Street	Nelson, C. Emory
	Jackson, 1019 First Street	Planck, R. M.
Mississippi	Mississippi City, P. O. Box 117	Pecoul, Ferdinand Anthony
Montana	Forsyth, Box 1064	Roberson, Carl
Nevada	Reno, University of Nevada	Sandorf, Irving Jesse
New Jersey	Bridgeton, 150 East Avenue	Nichols, Howard Leslie
	East Orange, 71 Leslie Street	Woodworth, Fred B.
	Jersey City, 169 Zabriskie Street	Weber, Walter
	Orange, 454 Conover Terrace	Kynor, Merrill W.
	Union City, 301–33rd Street	Masalkovics, Joseph Albert
New York	Brooklyn, 186 Hopkinson Avenue	Potter, Max
	Brooklyn, 164 Columbia Heights	Stinchfield, J. Maxwell
	Buffalo, 46 Riley Street	Oversmith, La Rue H.
	Buffalo, 605 Auburn Avenue	Perau, Frederic Henry
New York (cont'd.)	Buffalo, 62 Manden Street	Smith, Stanley C.
	Gardenville, 447 Potter Road	Felmet, Albert
	New York City, c/o Norton Lilly, 26 Beaver Street	Berry, Harold C.
	New York City, 140 West 16th Street	Castaneda, Santiago
	New York City, 118 East 103rd Street	Doubles, David J.

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	New York City, 41 West 86th Street	Feldstein, Martin A.
	New York City, c/o Norton Lilly, 26 Beaver	Francis, Philip
	Brooklyn, 326 Utica Avenue	Gottdenker, Martin
	New York City, R. C. A., 326 Broadway	O'Connor, John G.
	New York City, 195 Broadway	Reinken, Louis W.
	Rochester, 7 Edmonds Street	Steneri, Arthur John
	Schenectady, 848 Union Street	O'Neill, John P.
	Staten Island, USS *Acushnet*	Miller, Paul E.
North Carolina	Henderson	Woolard, E. W.
North Dakota	Fargo, 905-5th St., North, No. 19 Rust Apts	Cook, Tedd W.
Ohio	Cleveland, 7617 Myron Avenue	Crocus, John Stanley
	Cleveland, 3516 Storer Avenue	Irvine, Robert P.
	Marion, 177 East Center Street	Brown, David Ashton
	Zanesville, 606 Ridge Avenue	Bell, Lewis M.
Oklahoma	Oklahoma City, Oklahoma Gas & Electric Co	Bathe, C. E.
Oregon	Portland, 873 Belmont Street	McCargar, S. Harold
Pennsylvania	Allentown, 246 S. Madison Street	Bowman, Charles W.
	Allentown, 226 N. 6th Street	Haines, A. J. D.
	Allentown, 2014 Highland Avenue	Muthart, John A.
	Allentown, 949 Hamilton Street	Rauhofer, Frank
	Easton, 426 S. 21st Street	Clendaniel, John Edwin
	Easton, 18 N. 9th Street	Raesly, James B.
	Easton, 1815 Fairview Avenue	Weller, Everett Clare
	Philadelphia, 2209 South Chadwick Street	Johnson, Harmon
	Phialdelphia, 765 S. 10th Street	Mattia, Ralph F.
	Wilkinsburg, 215 Ross Avenue	Reynolds, C. C.
Rhode Island	Providence, 4 Pemberton Street	Brewster, O. H.
West Virgina	Charleston, 4 Maple Terrace	Moore, Thomas H.
Wisconsin	Milwaukee, 1667 Oakland Avenue	Hough, W. E.
	Milwaukee, 1405 Bremen Street	Strassman, Irving H.
British Guiana	Demerara, Georgetown, 61 Hadfield St	Tasker, Joseph Thomas
Chile	Valparaiso, Casilla 1653	Vierling, Gustav
England	Stockport, North Reddish, 446 Gorton Road	Howard, C. Alexander
India	Baroda, Babajipura	Dighe, K. S.
Ireland	60 Clifton Road, Bangor, Co. Down	Jamison, A.
Mexico	Puerto Mexico, Veracruz, Apartado, 86	Bourgeois, Allen B.

Elected to the Junior grade

Illinois	Champaign, 61 East Green Street	Pennington, D. J.
Indiana	Valparaiso, 712 Calumet Avenue	Clark, Edgar J.
	Valparaiso, 156 S. Franklin Street	Woodworth, Elwyn Crane
Massachusetts	Quincy, 189 Common Street	D'Alessandro, Genaro
Michigan	Royal Oak, 286 West Ten Mile Road	Alain, J. E.
New York	Brooklyn, 140 Vanderbilt Avenue	McGonigle, William J.
Oregon	Salem, 2405 Center Street	Poujade, Donald G.
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APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before March 1, 1929. These applicants will be considered by the Board of Direction at its March 6th meeting.

For Transfer to the Fellow grade

New York	New York City, 67 Broad Street	Pratt, Haraden

For Election to the Fellow grade

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For Transfer to the Member grade

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For Election to the Member grade

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Oklahoma	Ponca City, 604 Marland Drive	Wyckoff, Ralph D.
England	Cornwall, Bodmin, Radio Beam Station	Struthers, G. A.

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	Oakland, 1609 Grand Avenue	Wild, Sidney J.
	Pasadena, 230 South Hudson Avenue	Taylor, Kenneth
	San Francisco, 328 Custom House	Kellogg, Frank L.
	San Francisco, 9 Duncan Street	Pinkney, J. L.
	San Pedro, USS California, Box 17, c/o Postmaster	Hart, Clyde Armand
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	New Haven, 1523 Chapel Street	Rhoades, A. L.
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	Washington, U. S. Naval Research Laboratory	Wallace, James D.
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	Atlanta, Box 1593	McKinney, Joe H.
	Fort Benning, P.O. Box 615	Sosebee, Frank E.
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Part II

TECHNICAL PAPERS

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the anode is strongly negative, at about 100 volts, there is no longer any noticeable anode current. Gill and Morrell obtained oscillations with the anode completely insulated. Electrometric measurements of the potentials of the free anode showed it to be strongly negatively charged. When the *B* oscillations start its potential increases. The curves of oscillation intensity I_h with a strongly negative anode and with a free anode are almost identical (cf. curves 5 and 6 in Fig. 3).

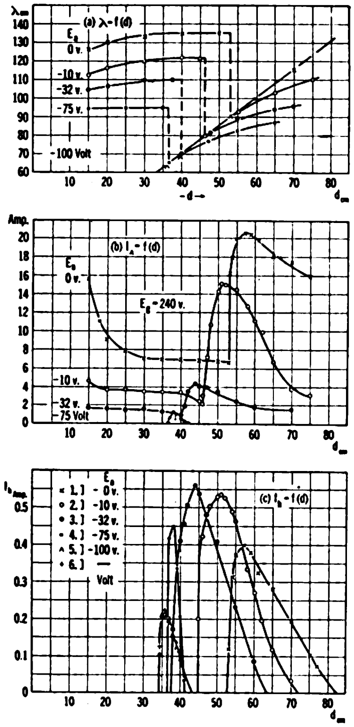

Fig. 3

The intensity of the *B* oscillations is considerable in comparison with type *A*; high-frequency currents up to 0.6 amperes occur in the thermo-element. The *A* oscillations could not be detected with the thermo-element even with the aid of a rather sensitive galvanometer. This shows that the external oscillation system is almost non-participating in the *A* oscillation process.

In the above case the maximum oscillation intensity was obtained at $E_p = -32$ volts, $d = 44$ cm, and $E_g = 240$ volts. The

current in the thermo-element was 0.56 ampere, the wavelength 78 cm. Accordingly it can be assumed that there is a most favorable wave for each grid potential, at which the intensity can be maximized by the proper choice of the anode potential and tuning.

3. The Transition

In the first experiments it seemed as if the Barkhausen-Kurz (*A* range) oscillations changed abruptly into the Gill and Morrell oscillations in the *B* region. More careful experiments, however, have shown that the *A* oscillations decrease slowly, and that the *B* oscillations appear suddenly with great intensity. Therefore,

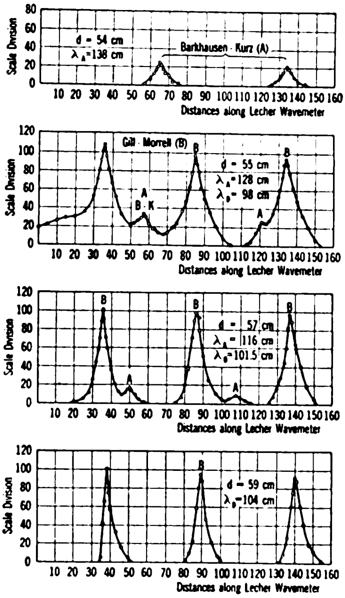

Fig. 4

in a narrow region both oscillations can be present simultaneously. Measurement of the differing frequencies by means of the Lecher system is generally impossible because the maxima of the Gill and Morrell oscillations mask the far weaker Barkhausen-Kurz oscillations. However, it was observed that the *A* frequencies no longer remained constant after the *B* oscillations started, but drifted upward a few per cent. Fig. 4 shows the conditions at the transition in a particularly clear case; here the intensity distribution on the Lecher system was recorded over

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several maxima for lengths d differing slightly. At $d=54$ cm only Barkhausen-Kurz electron oscillations are present; on displacing the bridge but one centimeter, however, the Gill and Morrell oscillations predominate. At $d=57$ cm, the A oscillations are weak and at the same time their wavelength has decreased from 138 to 101.5 cm. Finally, with $d=60$ cm, only the short-wave B oscillations of Gill and Morrell are present.

4. Influence of the External Oscillation System

For the production of the highest possible frequencies it is important first to generate "pure" electron oscillations of the A type which serve to initiate the shortest possible waves. According to the theory of Barkhausen-Kurz and the investigations of Scheibe this may be brought about by increasing the field strength between the electrodes. Secondly, the beginning of the Gill and Morrell oscillations is to be displaced as far as possible toward the short waves.

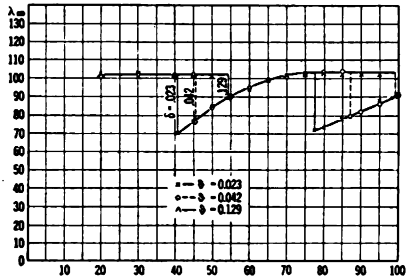

Fig. 5

Fig. 5 shows that the starting point of the Gill-Morrell oscillations is primarily determined by the character of the external oscillation system. The curves represent the effects of the high-frequency systems of different damping. This was done without shifting the tuning, by using materials with different conductivity for the parallel wires. As shown, the beginning of the Gill and Morrell oscillation is delayed by increasing the damping.

It is to be expected that a reduction of the damping must cause an extension of the B region. Experiments in this direction were not made, but this is explained by an observation given by Scheibe.[14]

[14] Scheibe. *Jahrb. d. Draht. Tel.*, 27, 1, 1926.

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In order to obtain Barkhausen-Kurz oscillations of greater intensity with parallel wires, Scheibe coupled two tubes together by their resonance circuits. Each of these tubes was able to set up independent oscillations, and he found that with external excitation by one, the other tube started to oscillate with 30–40 per cent lower potentials than was required for spontaneous oscillation. If we assume that the oscillations produced by Scheibe were of the Gill-Morrell type, as is probable on account of his arrangement, the extension of the oscillation range is easily explained as a reduction in damping by the external excitation which naturally acts in the opposite sense to the increase in damping previously described.

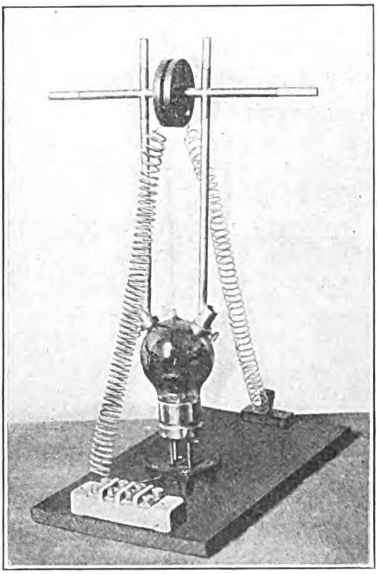

Fig. 6

During the course of the investigations a large number of widely different types of tubes were tried. Powerful electron oscillations of the *A* and *B* types were obtained with the French short-wave tube of the TMC type. This tube gave a 38-cm wave with a high-frequency current of 0.12 ampere. Fig. 6 shows a short-wave oscillator connected with this tube. To increase the radiation the tuning bridge has been extended on both sides to form a symmetrical antenna.

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5. Theoretical Observations

(a) Calculation of the Electron Vibration Frequency in the Alternating Field. Barkhausen-Kurz gave a formula for the calculation of the electron frequency under simplified conditions. If in the calculation we consider instead of the stationary retarding field an alternating field, we may expect theoretically a change in the electron motion and with it a change in the frequency. The complete period is given as the time spent by an electron in the grid-anode and grid-cathode spaces. It will be assumed that both spaces are equal so that the calculation need be carried out for one space only.

Assumptions:

(1) Anode and cathode potentials are zero.

(2) The grid is located midway between anode and cathode; that is, $d_a = d_k = d$ (see Fig. 7).

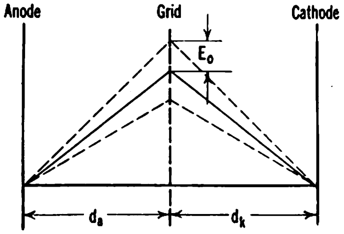

Fig. 7

(3) The electrodes are plane.

(4) Space charges are disregarded.

(5) The oscillation produces an alternating potential only at the grid.

Then the electric force in the inter-electrode space is:

$$E = \frac{E_g + E_0 \sin(\omega t + \phi)}{d}$$

and the equation for the motion of an electron is:

$$-e \cdot \frac{E_g + E_0 \sin(\omega t + \phi)}{d} = m \frac{d^2x}{dt^2} \cdot \qquad (1)$$

If we assume that the electron enters the inter-electrode space at time $t = 0$ with the velocity v_0 and that at this instant also $x = 0$, by double integration of (1) we get:

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$$x=-\frac{e}{m\cdot d}\cdot E_g\cdot\frac{t^2}{2}+v_0\cdot t+\frac{e}{m\cdot d}E_0\left(\frac{\sin(\omega t+\phi)}{\omega^2}-\frac{t\cdot\cos\phi}{\omega}-\frac{\sin\phi}{\omega^2}\right)\quad(2)$$

A simple solution of this equation for t is possible with the following assumptions:

(1) The frequency of the alternating field is determined by the electron frequency. Therefore, if we make $x=0$ in (2), we get the time t, which corresponds to a half period; this is:

$$\omega\cdot t=\frac{\omega T}{2}=\pi$$

(2) At the time $t=0$, that is, at the instant of the passage of the electron through its meshes, the grid receives a negative charge, as most of the electrons strike the grid. At this instant, therefore, E_0 has its maximum negative value, that is, ϕ must be equal to $-\pi/2$.

Then the duration of a semi-period is·

$$\tau=\frac{v_0}{\frac{e}{m\cdot d}\left(\frac{E_g}{2}-\frac{2E_0}{\pi^2}\right)}\quad(3)$$

In which

$$v_0=\sqrt{\frac{e}{m}2(E_g-E_0)}$$

If in (3) we substitute the value of v_0 and of the potential in volts, we get the wavelength

$$\lambda=c\cdot 2\cdot\tau$$

$$\lambda=\frac{4000\sqrt{E_g-E_0}\,d}{E_g-\frac{4E_0}{\pi^2}}\quad(4)$$

Concerning this equation (4) we notice first that when $E_0=0$ it changes into the equation given by Barkhausen-Kurz. It is only necessary to replace the electrode space d by the "anode diameter" d_a which is equal to $4\cdot d$, *ex hypothesi*.

There is a shortening of the wavelength due to the alternating potential, E_0, as is seen from the following table:

$E_g=500$ volts, $d=0.5$ cm

E_0	λ
0	89.5 cm
100	86.9
200	82.5
300	74.6
400	58.8

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If in (2) we substitute various values for t starting from $\omega t = \pi$ we get the curve of Fig. 8 for the path described by the electron. In this figure different values of E_0, alternating potential, are illustrated. We see how the excursions of the electron continually diminish with increasing E_0 and the electrons always return sooner to the grid. Due to the simplifications, the curves give the course of the electron motion only qualitatively, and a hopeful comparison with experimental values cannot be attempted.

(b) Mechanism of the Gill and Morrell Oscillations. On the basis of the above observations the mechanism of the Gill and Morrell oscillations can be explained as follows:

Assume that the oscillation circuit tuning is changed from shorter to longer waves. At first its natural wavelength is far below the excitation frequency of the Barkhausen-Kurz oscilla-

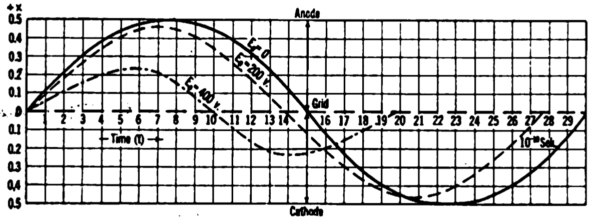

Fig. 8

tions; hence the alternating potential between the electrodes is very small and we have "pure" electron oscillations in the sense of Barkhausen and Kurz. (See Fig. 8, $E_0 = 0$.) As the oscillation circuit is tuned to the exciting wave, the alternating potential at the electrodes rises, corresponding to the resonance curves, until a certain potential, designated as the "ignition potential," is reached.

Now the alternating field superposed upon the stationary retarding fields causes a shortening of the exciting wave so that the attainment of resonance is accelerated; the result is an additional strengthening of the alternating field and so on, that is, a building-up process takes place which rapidly affects both the alternating potential and the frequency. The final stable state is reached when the frequency of the electron vibrations coincides with the proper frequency of the oscillating system, for an increase in the excitation frequency above resonance would cause a weakening of the alternating field. From this it is evident

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that the frequency of the Gill and Morrell oscillations is determined only by the natural wavelength of the tuning system.

On the other hand, if the natural wavelength of the oscillation circuit is above the Barkhausen-Kurz wavelength, an alternating field also arises on approaching resonance, but as this reduces the exciting wave its action is opposite to the above and causes a displacement from resonance. A building-up process is therefore discouraged. Only on approaching the exciting wavelength from longer natural wavelengths are the Barkhausen-Kurz waves gradually shortened, and thus we get the explanation of the course of the frequency curves in Fig. 2b as well as Fig. 3a, with its abrupt start on one side and its gradual transition on the other.

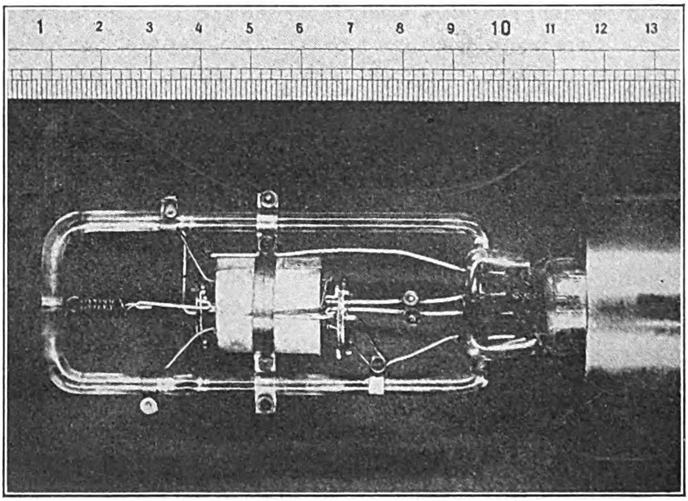

Fig. 9

It is proper to find analogies between the above building-up process and the building-up of oscillations in regenerative circuits. Therefore it is not to be wondered at that the well known "pulling" (ziehen) of the usual transmitter was also observed in the oscillations present here. The starting point of the Gill and Morrell oscillations is not determined uniquely, but is displaced somewhat according to whether the tuning of the system is varied from longer to shorter waves or the reverse. For this reason the data of the curves in these illustrations were all taken in the same direction, namely from shorter to longer waves.

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Summarizing the results of Part I it may be stated that two different kinds of electron oscillations occur with the retarding field connection of Barkhausen-Kurz, those of a pure Barkhausen-Kurz type, in which the frequency is determined by the stationary interelectrode fields, and those of Gill and Morrell whose frequency agrees exactly with the natural frequency of the tuning system. Between these types of oscillations there is a region of steady transition, in which there is a departure from the true wavelength of the oscillation circuit and a gradual transition toward the Barkhausen-Kurz frequency. Therefore if "pure" oscillations of one or the other form are desired, the external system must be carefully and properly tuned. Carelessness in this tuning process may be suggested as the cause of many disagreements between different authors. In this connection it must be observed that the leads or tube electrodes may themselves comprise an oscillation system and be capable of producing Gill and Morrell oscillations.

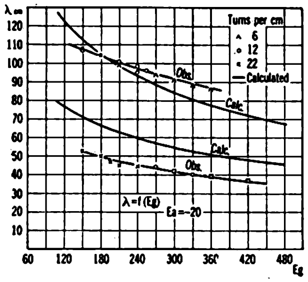

Fig. 10

II. Electron Oscillations of Similar Type of Different Frequencies

1. Shorter and Longer Barkhausen-Kurz Oscillations

Scheibe[4] has already reported Barkhausen-Kurz oscillations of two different frequency ranges. In the course of my own investigations I likewise observed a higher frequency wave range with the lower range, but could not reproduce it with certainty. I later substituted for the Schott tube a specially constructed electrode system supported by a small glass frame which by means of a ground joint could be placed in a glass vessel connected with a vacuum pump.

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For the generation of highest frequencies it is naturally of the greatest importance to ascertain the conditions under which the higher frequency range of the Barkhausen-Kurz oscillations occurs. On account of the large number of frequency-determining factors the variables must be restricted as much as possible and for this reason the radial dimensions of the tube, that is, the grid and anode diameters, were first maintained constant. The only variable factor which remained was the pitch of the grid.

As the Schott tubes have an average of 8–10 turns per cm of grid length, a grid with 15 turns per cm was made and inserted in the anode cylinder of a Schott tube. The study of this system showed that under the same operating conditions strong "shorter" waves were readily obtained, whereas even at higher grid potentials no "longer" Barkhausen-Kurz oscillations could be detected. For more systematic investigation of the distribution of the two frequencies over the entire oscillation range under different operating conditions, the same grid system was successively provided with windings of 8 different pitches and with different grid potentials. In order to exclude Gill and Morrell oscillations with certainty, the electrodes, as can be seen in Fig. 9, were bridged on both sides by two blocking condensers and the external oscillation system, or leads, are carried off in this way.

The results of the investigation are shown in Fig. 10, in which both frequencies are given as functions of the grid potential. With 6 windings per cm the "longer" waves are present exclusively; with 12 windings per cm the short waves appear at $E_g = 260$; with 22 windings per cm these start at $E_g = 150$ volts. Then the longer waves appear with the shorter up to 180 volts. It is clear from the measurements that the shorter waves start the more readily the smaller the pitch of the grid helix. If we compute the ratio of the wavelengths, the resulting values differ from 2. Thus, as Scheibe has emphasized, it is demonstrated that the shorter waves cannot be regarded as overtones of the longer.

2. Mechanism of the Shorter and Longer Barkhausen-Kurz Oscillations

The observation that the permeability of the grid is determinative for the appearance of the longer or shorter Barkhausen-Kurz oscillation leads to the following explanation: In their theory Barkhausen and Kurz assume that the electrons swing

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about the grid; this view is fortified by the experimental agreement. In Fig. 11 on the left there is shown the motion of an electron which swings twice about the grid. As the anode has a negative potential the electron reverses in front of it at the position of the surface of zero potential.

A large part of the electrons emitted by the cathode, of course, do not penetrate the grid after their first journey through the grid-cathode space, but strike the grid wires. These electrons impart all their energy to the grid, and need not be considered in the generation of oscillations. With the electrons which are reversed in the grid-anode space but which do not pass through the grid the second time, however, the case is quite different. On account of their vibratory motion in the retarding field of the anode, they are capable of supplying energy for the production of oscillations. The frequency of these oscillations is determined

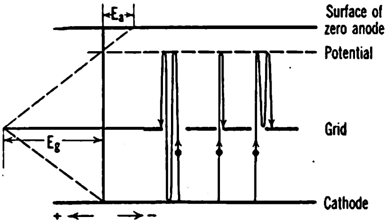

Fig. 11

only by the stay of the electrons in the grid-anode space and must therefore be higher than the electron oscillations in the anode-cathode space itself. At the right of Fig. 11 is shown the motion of an electron which moves to and fro once in the grid-anode space. It must also be remembered that the electrons can swing to and fro several times in the grid-anode space, chiefly under the influence of the alternating potentials at the grid. We are then dealing with Gill and Morrell oscillations, which will be discussed later.

On the basis of this picture of the mechanism of the higher frequency electron oscillations, it is comprehensible that the permeability of the grid influences the appearance of one frequency range or the other. The narrower the mesh of the grid the lower will be the percentage of electrons that pass through it until finally no vibratory motion at all takes place about the grid, it being then entirely confined to the grid-anode space. This could be shown experimentally.

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Scheibe has already considered the stay of the electrons in the grid-anode and in the grid-cathode spaces separately. His formula for the calculation of the wavelength permits the direct calculation of the two partial frequencies separately. This calculation is carried out for the system used and the result is shown in Fig. 10. It is seen that the longer waves show a considerable deviation from the experimental values; neither is a quantitative agreement of the shorter waves to be expected, so that a proof could not be adduced from this.

3. Experimental Tests

The wavelength of the Barkhausen-Kurz oscillations involving the entire interelectrode space will be designated by λ_{AC}; as has already been explained they consist of two partial oscillations, namely the electron motion in the grid-anode space whose wavelength is λ_{GA} and in the grid-cathode space of wavelength λ_{GC}, so that

$$\lambda_{AC}=\lambda_{GA}+\lambda_{GC}$$

Only the waves λ_{AC} and λ_{GA} can be excited; λ_{GC} cannot appear physically, at least not in a normal three-electrode tube. It can be determined indirectly experimentally from the difference between the longer and the shorter waves, for according to the above equation:

$$\lambda_{GC}=\lambda_{AC}-\lambda_{GA}$$

Constant operating conditions being assumed, λ_{GC} must depend entirely on the dimensions of the grid-cathode space. If the dimensions of the grid-anode space are changed by increasing the anode diameter, only a change in λ_{AC} and λ_{GA} can be expected, while their difference, λ_{GC}, must remain constant.

I have previously discussed the dependence of both waves upon the grid potential (Fig. 10) for an anode diameter of 18 mm. With the same grid inserted in an anode with a diameter of 26 mm, and with a variation of the grid pitch, the longer and shorter waves are found over the same potential range. The new system is shown in Fig. 12. It is also shunted by two blocking condensers so that only "pure" electron oscillations can occur. The curves in Fig. 13 show the frequency; in spite of the great change in grid winding, both waves could be obtained over a small potential range. At low grid potentials the frequency deviates from the direction expected from theory, increasing

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with the potential; only above 180 volts does the course correspond with the theory. This abnormal action is probably due to the fact that at low potentials the space charge is accentuated

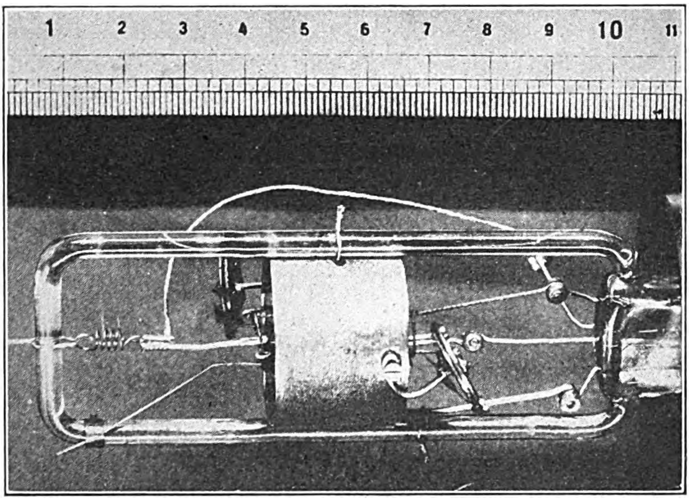

Fig. 12

and its effect on the frequency, especially in the large grid-anode space, may become of importance.

The difference in both curves gives λ_{GC} as a function of the operating potentials for each system. The course of this curve

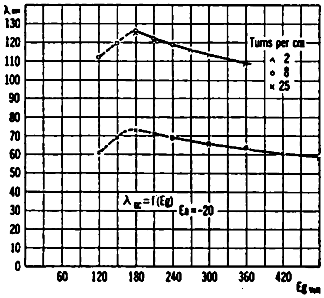

Fig. 13

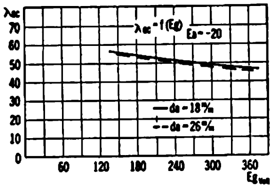

Fig. 14

is shown in Fig. 14, the solid curve for the 18 mm system, and the dotted curve for the 26 mm system. The close correspondence of the grid-cathode waves of the two systems furnishes excellent support for the above view of the mechanism of the shorter electron waves.

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4. Gill-Morrell Oscillations of Higher Frequency

If the shorter electron oscillations are a special form of the Barkhausen-Kurz oscillations one may expect that, like the longer waves, they are affected by the alternating field between the electrodes. This follows from the theoretical observations of Section 5 of Part I which hold as well for the grid-anode space as for the whole anode-cathode space.

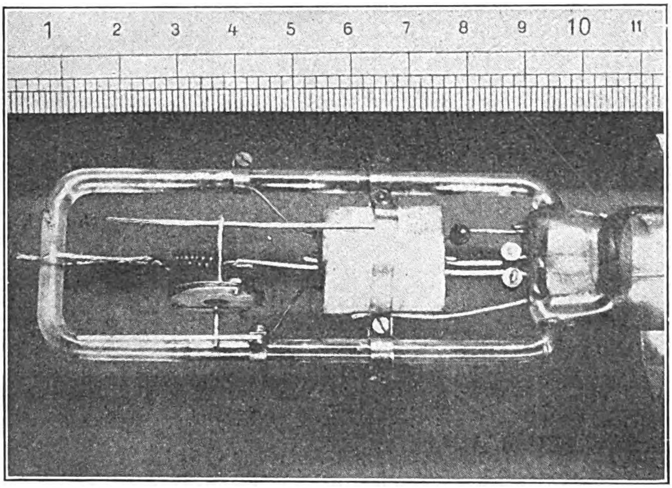

Fig. 15

In the effort to produce the highest possible frequencies, a system was constructed in which the electrodes were excited in antiresonance, and in which the oscillation-circuit was partially inside the tube and partially outside. If one oscillation node is fixed by the position of the internal condenser (see Fig. 15), the

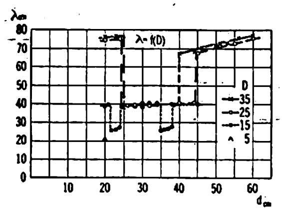

Fig. 16

other can be adjusted by shifting the bridge on the external parallel wires. Fig. 16 shows the frequency curves for constant operating potentials of 360 and −20 volts. *D*, the length of the

wire system within the tube, is taken as a parameter. Gill-Morrell oscillations of lower or higher frequency are set up depending upon the tuning condition. With D equal to 5 mm, a wave of 21.7 cm can be produced whose intensity is so great that measurement of wavelength could be made at a distance of 2 meters.

In order to produce still shorter waves a system of low damping was constructed in which only the half-wave oscillation was permitted. As shown in Fig. 17, for this purpose grid and anode are equipped *on both sides* with longitudinal wires along which capacity bridges can be shifted. The entire oscillation system is thus within the tube.

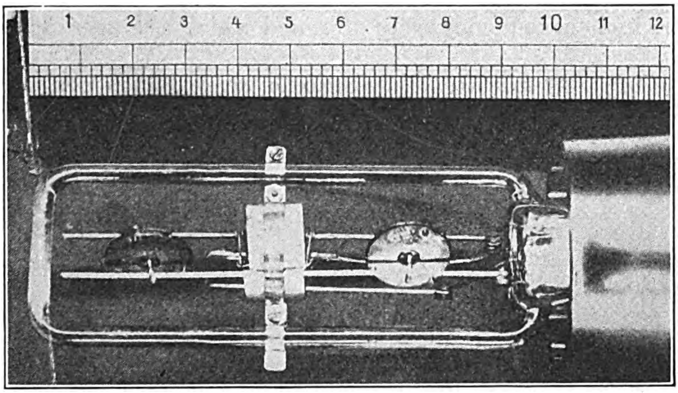

Fig. 17

With this system two frequency ranges could be obtained by varying the operating conditions, first the Barkhausen-Kurz higher-frequency oscillations (as a comparison with earlier measurements showed) and second, the Gill-Morrell higher

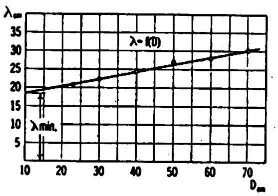

Fig. 18

frequency oscillation, whose wavelength is given by the distance of the two condensers in the tube. If the Gill-Morrell wave is plotted as a function of this distance, we get the calibration curve of the oscillator (Fig. 18).

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The smallest adjustable length, D, was 23 mm; for this there was obtained a 20.8-cm wave. Extrapolating the curve to smaller D, which is permissible on account of its linearity, we find for $D=10$ mm, the length of the electrodes themselves, a wavelength of 18.5 cm; *this represents the shortest wave that can be obtained with the present system.*

During all previous investigations the radial dimensions of the electrodes were kept constant. These are potential additional means for reducing the Barkhausen-Kurz frequency. The system shown in Fig. 19 was therefore constructed, in which the anode was reduced to a diameter of 11 mm, and the grid to a section of 3×3 mm. With this system Barkhausen-Kurz oscillations of the higher frequency were obtained which averaged an *octave*

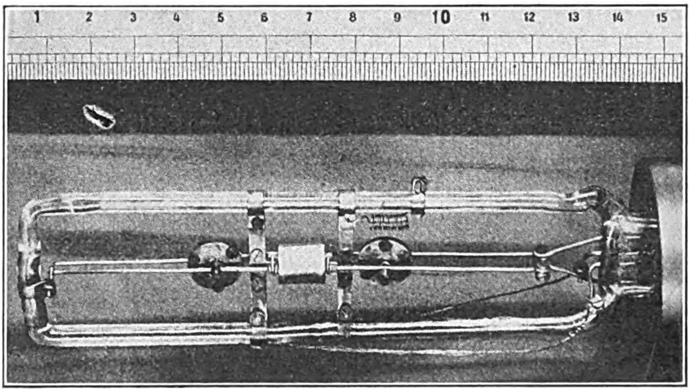

Fig. 19

higher than the previous system. By suitable tuning of the oscillation system Gill-Morrell oscillations of higher and lower frequency could be produced. The shortest tuning-system length was fixed by the fastening-pieces of the heating filament at 36 mm; for this there was a natural wave of 22.5 cm. With a grid potential of 360 volts Barkhausen-Kurz oscillations of 21.4 cm were still obtained; accordingly, with a proper tuning system Gill-Morrell oscillations down to 15 cm might be obtained with safety.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

A CONVENIENT METHOD FOR REFERRING SECONDARY FREQUENCY STANDARDS TO A STANDARD TIME INTERVAL*

By

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(Research Laboratory, General Radio Co., Cambridge, Mass.)

Summary—*A method is described for obtaining a convenient low frequency from a high-frequency standard by means of harmonic control of distorted wave oscillators (multi-vibrators). In the equipment described two such oscillators are employed, for convenience in arriving at a harmonic series of frequencies, based on 10-kc fundamental, as well as a frequency of 1 kc for operation of the clock motor, from a standard frequency of 50 kc. The conclusions based upon an experimental investigation of harmonic control of the distorted wave oscillators are given. Representative curves of the variations in frequency of the 50-kc standard for 10-day intervals are shown.*

1. Introduction

In recent years the technique of frequency standardization has been refined and improved through the development of oscillating systems which are so nearly invariant that their average frequency is practically independent of the time interval for which the average is determined. Such a system permits the determination of frequency, by counting successive cycles, to a degree of precision which is usually limited only by the patience of the observer. Extensive use of this "absolute" method of frequency measurement has suggested the concept of the earth's rotational frequency, or a time interval defined thereby, as a fundamental standard for frequencies of all magnitudes.

Practical methods for maintaining various types of oscillators as the intermediate links between the desired frequency and a standard time interval have been described in the literature. It is the purpose of the present paper to call attention to an application of the same general principle to a specific problem of laboratory calibration. The results described have a general significance insofar as they relate to the establishment of a harmonic series of standard frequencies.

* Dewey decimal classification: R210. Original manuscript received by the Institute, November 15, 1928.

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At the present time there are approximately 600 radio broadcast stations operating more or less continuously in the United States. The transmitters of these stations have an irregular geographical distribution and to each transmitter is assigned as a carrier one of the 96 decimal frequencies between and including the limits of 1500-550 kc. The universal problem of interference is not solved by accurate maintenance of the carrier frequencies on their assigned values, but such maintenance is obviously a primary essential. Although piezo-electric oscillators, or their equivalent, will ultimately be incorporated as integral frequency control elements of all broadcast transmitters, such arrangements are not now in general use. Small portable piezo oscillators are widely employed in the broadcasting stations as comparison standards, by means of which their carrier frequencies may be occasionally corrected. The problem of accurate adjustment of quantities of these control and comparison standards to the even decimal frequencies employed by these stations has suggested the calibration method which we are to describe.

2. Experimental Method

The general technique of "direct" calibration from a single-frequency source, which in turn is compared with the earth's rotational frequency, is based upon an assumption which may be stated as follows: if oscillations of constant periodicity occur in two independent systems and either system be so adjusted that the fundamental constituent of its oscillation coincides constantly with a harmonic constituent of the oscillation in the other system, then the fundamental frequency of either system is uniquely determined in terms of the other. The propriety of this assumption will not be discussed here although other investigators have given attention to specific phases of it, as for instance by examining the validity of identifying with integral harmonic ratios the various emission frequencies from a multivibrator[1]. If, therefore, our fundamental oscillation occurs at a low frequency and is associated with a substantial series of real harmonics, the physical comparison of an unknown frequency with the frequency of one of these harmonics is considered to be an absolute comparison with the fundamental. Since the ultimate standard of frequency is very low indeed compared with what

[1] D. W. Dye, "A Self-Contained Harmonic Wavemeter", *Phil. Trans. Roy. Soc.* of **224**, A, London, 279.

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we normally have to measure, it is customary to employ the intermediate source of oscillations previously mentioned as the basis of the harmonic series from which calibrations are actually made.

An electro-mechanical oscillator such as a tuning fork has been extensively used as an intermediate source. A tube-driven fork can be coupled directly to the counting mechanism, but the fundamental frequency is inconveniently remote from the frequencies at which calibrations are required in the present case; also the fork is subject to the familiar limitations arising from its damping and its temperature coefficient. The use of a piezo-electric oscillator as the intermediate source has been described in an important paper read before the Washington meeting of U.R.S.I. in October, 1927.[2] The quartz oscillator exhibits small perturbations under reasonable operating conditions, and permits the establishment of a harmonic series from a fundamental of higher frequency. It requires, however, a very stable and reliable electrical system for converting its fundamental frequency into a lower frequency suitable for operating the counting mechanism.

In the system to be discussed in the present paper, a piezo-electric oscillator is employed as the intermediate source. But for the step-down frequency converter, a system describing "relaxation oscillations"[3] is adopted because it has been found to offer many advantages for the specific problem in hand. An intermediate frequency of 50 kc is chosen for convenience. The relaxation oscillator can be adjusted to yield a highly stable fundamental at 1 kc or even less, directly controlled by the 50-kc source, which operates a small impulse motor, running in air, for the counting mechanism. Furthermore, the controlled relaxation oscillator, if properly designed, is an extremely simple piece of equipment.

Two general types of relaxation oscillator have been suggested for use as step-down frequency converters. The simplest is a neon discharge tube, used in virtue of its so-called cut-off characteristic to sustain oscillations in a circuit comprising a direct-current source, a resistance, and a condenser. The oscil-

[2] J. W. Horton and W. A. Marrison, "Precision Determination of Frequency." Paper read at meeting of the International Union of Scientific Radiotelegraphy, Washington, D. C., Oct. 13, 1927. Proc. I.R.E., **16**, 137; February, 1928.

[3] Balth. van der Pol, Jr., "On Relaxation Oscillations," *Phil. Mag.*, 978, Nov., 1926, also *Zeits. f. Hochfreq. Technik*, **29**, 114, 1927.

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lations of such a circuit are rich in harmonics and if a second oscillation of a higher frequency is injected into the circuit, the relaxation oscillation may assume a frequency which is an integral submultiple of the injected frequency. It is possible to maintain an oscillation in such a circuit whose fiftieth harmonic coincides with the injected frequency. But the behavior of this control effect on such high orders is relatively unsatisfactory, as minute variations in either the injected voltage or the direct voltage are likely to destroy the control. The other type of oscillator is the Abraham-Bloch multivibrator. Doubtless there are many other forms of relaxation oscillator which present similar characteristics, but the modified multivibrator has proved so satisfactory that no others have been investigated.

The multivibrator consists essentially of an aperiodic circuit in which oscillations of irregular waveform are sustained by a triode excited by a second triode which provides the proper phase relation for maintenance. It was first used as a rich source of harmonics, in which the fundamental constituent is determined by and probably maintains a constant phase relation with a low-frequency injected oscillation. In its usual form it therefore constitutes a frequency multiplier or step-up frequency converter suitable for comparing high frequencies with a low-frequency standard. Some time ago one of the present authors[4] suggested the possibility of controlling a multivibrator at a fundamental frequency which is an integral submultiple of the injected control frequency, thus making it a step-down frequency converter. In this case the control oscillation coincides with a harmonic of the multivibrator. When this condition is established experimentally, the multivibrator is definitely controlled by the injected high-frequency oscillation, and the multivibrator oscillation has certain characteristics which distinguish this "controlled" state of oscillation. Since this system appears to be well adapted for driving a counting mechanism at frequencies as low as one one-hundredth of the constant frequency source, we shall examine in detail its behavior in the role of a controlled submultiple generator.

[4] J. K. Clapp, "Universal Frequency Standardization from a Single Frequency Source," *Jour.* Opt. Soc. Amer. and Rev. Scien. Instr., 15. 25, July, 1927.

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3. Harmonic Control of Multivibrator

No adequate mathematical analysis of the action of the multivibrator has been given. In the practical case large voltage amplitudes are encountered, carrying the operating points of the vacuum tubes far beyond the region of the characteristic which might be treated under limited series solutions for small amplitude oscillations such as van der Pol's.[3] But large amplitudes are required to obtain an extended series of harmonic frequencies of useful amplitude. It thus appears that the only feasible investigation of multivibrator oscillations, meeting our physical requirements, demands sufficiently complete experimentation to permit valid generalizations to be drawn.

The basic assumption of integral ratios between the harmonics and fundamental *emitted* by a multivibrator allows frequency comparisons to be made, but does not necessarily imply that the controlled multivibrator has a harmonic frequency, of value equal to the frequency of the controlling source, and of invariant phase with respect to this source. A regular, or irregular, variation in phase might conceivably exist.

We define the "controlled" state for the multivibrator as that state of operation in which one of the multivibrator harmonics does not change phase with respect to the fundamental frequency of the controlling source by more than 360 electrical degrees in any arbitrary time interval. Then, by the assumption of integral harmonic frequencies, the fundamental frequency of the multivibrator is an exact submultiple of the controlling frequency. We shall next examine the experimental conditions for permanently establishing this controlled state on the desired harmonic order.

All the experiments have been carried out with no inductance, save that of strays, in the circuits of the multivibrators because it seemed logical to reduce the electrical inertia to a minimum when searching for the most stable and unambiguous control condition.

Preliminary studies indicated that the control stability in any multivibrator was increased by the injection of the control voltage into the common plate lead of the multivibrator tubes, as compared with injection into the circuits of either tube individually. Such increased stability is obtained, however, only if the multivibrator tube circuits are unsymmetrical so that an appreciable resultant plate current exists in the common lead.

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The use of the unsymmetrical multivibrator in this connection resulted from earlier work by one of the authors with the symmetrical arrangements. Dissymmetry was first attained through the use of tubes of different type, having widely differing filament emissions. Later work showed the same result was obtained with tubes of the same type, if the circuit constants were made dissymmetrical, as by the use of resistances 20 to 50 times as great in one tube plate circuit as in the other.

As a result of this dissymmetry, one tube operates as a "throttle," the operating point sweeping across the entire characteristic from saturation to cut-off, while the other operates substantially as a normal amplifier. This type of operation, no matter how attained, is characteristically favorable for stable control.

A tube is employed between the controlling source and the multivibrator which does not function primarily as an amplifier, but as a one-way relay, whose purpose is to isolate the standard source from reactions due to the multivibrator. This is not only generally desirable, but it is a necessity if the standard is to maintain its frequency with extreme constancy. An ordinary vacuum-tube amplifier circuit is not sufficient when the controlling frequency is as high as 50 kc. Either a tetrode, utilized as a "screened grid" amplifier, or a carefully neutralized triode, must be employed.

In this work the choice of the standard was made in favor of a very heavy quartz bar whose frequency lay close to 50 kc. For convenience, only, the order of frequency division was taken as 50, so that the controlled multivibrator fundamental would be close to 1 kc. The impulse motor and clock train were designed to keep correct time when the frequency applied was 1,000 cycles per second, so that with this division the timing of the high-frequency source is reduced to observations of small time errors indicated by the clock.

The schematic circuit arrangement for a single controlled multivibrator is given in Fig. 1. In the absence of the tetrode and the resistance R, the uncontrolled multivibrator remains. In this work the resistances were fixed and variation of the fundamental frequency of the multivibrator attained by simultaneous variation of the capacities C. In the uncontrolled state, variation of these capacities produces a corresponding smooth variation of fundamental frequency of the multivibrator.

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The application of a controlling voltage, e_c, to the grid of the isolating tube results in the injection of a voltage of the same frequency into the multivibrator circuit. Provided that the voltage is sufficiently great, that is, greater than a few hundredths of a volt, variation of the condensers C no longer produces a smooth and continuous change of fundamental frequency of the multivibrator. For certain separated ranges of values for C, the fundamental frequency of the multivibrator assumes discreet values, these being submultiples of the control frequency and invariant over the particular range of C. Between the ranges of C at which these discreet frequencies exist there are "transition" ranges in which no stable value exists for the fundamental frequency of the multivibrator. In these transition ranges the multivibrator attempts to oscillate in the controlled and uncontrolled states simultaneously, or, if the control voltage be

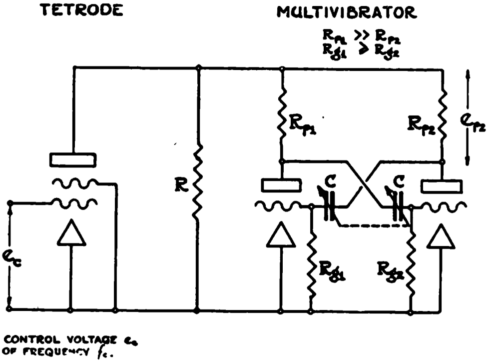

Fig. 1—Multivibrator of Fundamental Frequency f_c/n, where n is order number of harmonic at which control takes place.

sufficiently great, in adjacent controlled states. Observed by oscillograph, it is seen that in the latter case the fundamental oscillation of the multivibrator consists of a number of cycles of one discreet frequency and then one or more cycles at another discreet frequency. The oscillations alternate between the two frequencies in an irregular manner, there being no definite number of cycles in sequence of either of the two frequencies.

With small values of controlling voltage the operation is generally such that in the transition ranges of capacity, between those at which discreet and invariant values of fundamental frequency are obtained, the multivibrator oscillates in the uncontrolled condition. The regions in which control is attained

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are well defined and the controlled state is fairly stable. This mode of operation may be termed the "under-voltaged controlled state." If the control voltage be increased in magnitude, the transition ranges of capacity over which uncontrolled or unstable operation is obtained, are reduced progressively as the control voltage magnitude is raised, until finally the multivibrator fundamental frequency changes abruptly from one mode to the next upon varying the capacities through a very small transition range. This state of control may be termed the "over-voltaged controlled state" and represents that condition of operation which is believed to be the most stable and most desirable. In this condition, variation of capacity from maximum to minimum results in the production of a series of discreet fundamental frequencies for the multivibrator, the number of such frequencies which may be obtained being dependent upon the range of variation of capacity and upon the order of control which has been set up.

If the control voltage be increased beyond the point at which the transition ranges first shrink substantially to zero an effect which may be termed "drawing" is produced. This drawing effect consists of a progressive increase in the multivibrator fundamental frequency through discreet steps representing the successive harmonic control orders, as the amplitude of the control voltage is increased. In other words, the fundamental frequency is drawn toward the control frequency as the control influence is made stronger. If a particular control order is desired this drawing effect must be compensated by an increase in the multivibrator capacity or resistance. But this in turn tends to move the time constant of the multivibrator circuit away from the fundamental frequency at which it is being forced to oscillate, which ultimately reduces the control stability on its own account. The net result of these effects is the appearance of a fairly definite *optimum value* of control voltage for any given harmonic order.

We come now to somewhat more generalized conclusions concerning the dependence of the control range, the control order and the wave-form of the multivibrator oscillations upon variations of control voltage magnitude, variations in multivibrator fundamental frequency as produced by *symmetrical* variations in the multivibrator circuit constants, and by variations in the *degree of dissymmetry* of the multivibrator circuit pro-

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duced by a non-symmetrical variation of the circuit constants. These considerations are best indicated by an outline:

1.	Variations in control voltage, e_c	Small e_c	Limited control range, though control is stable within range. Waveform of same type for various orders of control.
		Large e_c.	Maximum control range; further increases of control voltage produce changes in order of control by "drawing," with narrowing of control range if e_c be made large enough.
2.	Variations in multivibrator fundamental frequency due to symmetrical changes in circuit constants. (C, C, varied).		Control order changes in successive steps to lower orders as fundamental frequency of multivibrator is raised toward control frequency. Control range narrow with small e_c; control range wide with large e_c. Waveform of same type for the various contro orders.
3.	Variations in degree of dissymmetry of multivibrator circuit. (R_{p2} only varied)		Control order remains constant. Waveform alters in discreet steps to forms in which the "positive" and "negative" portions of the wave are in *integral ratios of time*, the total time remaining constant.

Based upon the experimental evidence, an important generalization may be drawn. *Symmetrical changes* in *circuit constants* produce effects similar to *symmetrical changes* in *control voltage magnitude.* (Changes in control voltage are inherently symmetrical in the arrangement used, since the same voltage is injected in the plate circuits of both tubes.) These effects are characterized by control at successive orders with no major changes in the waveform. Changes in the *degree of dissymmetry* of the circuit result in *discreet changes of waveform,* the control order usually remaining fixed. These changes of waveform are characterized by discreet changes in the ratio of time of "negative" to time of "positive" portion of cycle, the total time remaining constant.

It is worthy of note that under no operating condition was the stability of control improved by distorting the waveform of the control voltage to produce harmonics which might "lock in" with the higher harmonics of the multivibrator. This appears to be due to the fact that the effect which we term "control" occurs at a certain definite phase relation between the control wave and the corresponding multivibrator harmonic, and the phase differences between the higher harmonics of both sources, although invariant, may either aid or oppose the control which is already established.

The oscillograms of Fig. 2 indicate the phenomena observed. These oscillograms are line drawings of experimental observations of the plate current in that multivibrator tube which operates through the low external plate resistance. This current has in all cases the same general waveform as the relatively minute

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plate current in the throttle tube and also the current in the common plate circuit which includes the control tube.

Diagram A indicates the changes resulting from variations of the magnitude of the control voltage, all other factors remaining constant. The waveform remains essentially the same. As the control voltage is raised, the fundamental frequency of the multivibrator is also raised, being "drawn" toward the frequency of the controlling voltage. These observations give a reasonable explanation of the fact that in a symmetrical multivibrator a strong preference is evidenced for control to take place at even ratios, because of the symmetry of the "positive" and "negative"

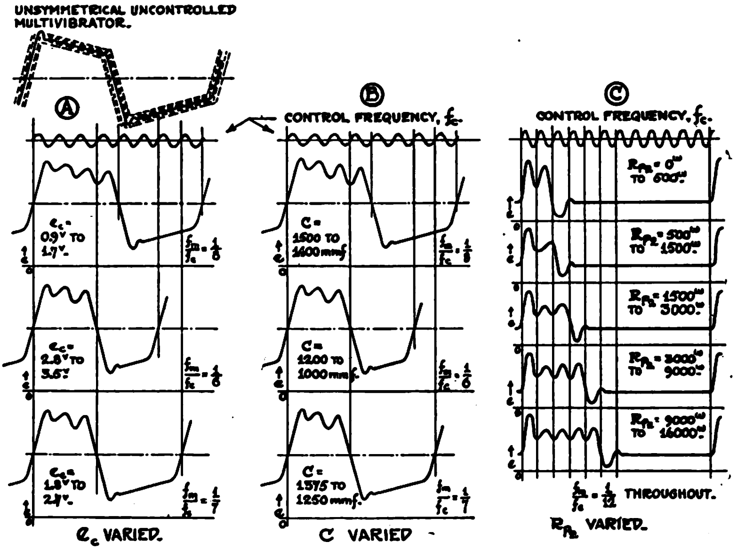

Fig. 2—Waveform and Frequency of Controlled Multivibrator, as shown by e_{p2} (Fig. 1), as functions of e_c, Cm, and R_{p2} as variables.

portions of the wave. With the dissymmetrical multivibrator such a preference does not appear to exist, which is explainable on the basis that the inherently unsymmetrical waveform readily adjusts itself so that one portion of the wave occupies a time interval equal to that of an odd number of cycles of the controlling frequency, while the other portion of the wave occupies a time interval corresponding to an even number of cycles of the control frequency. These conditions are illustrated in particular by the lower diagram for the frequency ratio of 1/7. Here the "positive" portion of the wave occupies a time interval of 3

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cycles of the control frequency, while the "negative" portion occupies a time interval of 4 cycles. If the waveform is more nearly symmetrical, the multivibrator must be forced to operate dissymmetrically, by the controlling influence, in order that odd frequency ratios may be obtained, consequently requiring a greater controlling influence than in the unsymmetrical case.

In diagram B are shown sketches of the waveforms resulting from variation, in a symmetrical manner, of the capacities C of the multivibrator circuit, all other factors remaining constant. The waveform is seen to remain essentially constant but the frequency assumes three discreet values, bearing the ratios 1/6, 1/7, and 1/8 to the controlling frequency. The most stable adjustment of capacities is at a value slightly below the middle of the control range.

These two diagrams, 2A and 2B, illustrate clearly the equivalent effects of symmetrical changes of controlling voltage and of circuit constants upon the waveform and frequency of the multivibrator oscillations.

Diagram 2C indicates the changes in waveform as the lower of the two plate circuit resistances of the multivibrator is varied. In this case R_{p1} was 400,000 ohms, R_{p2} variable between 0 and 18,000 ohms, and R was 10,000 ohms. The multivibrator remained in control at all values of R_{p2}, with a fundamental frequency which was one-twelfth of the controlling frequency. The successive discreet waveforms were obtained over considerable ranges of values for R_{p2}, as indicated by the marginal notations in the figure. The integral ratios of the times for the two portions of the cycle are clearly indicated, as well as the manner in which the serrations due to the controlling frequency shift by single units as R_{p2} was continuously varied.

The assembly finally employed consists of two multivibrators in cascade, the choice of ratios being determined by the available standard frequency, the normal frequency of the clock motor and the fact that a series of harmonics based on 10-kc fundamental frequency was desired for calibration purposes. While frequency ratios with a single multivibrator of the order of 50 may be obtained with stable operation, it would have been necessary to utilize a second multivibrator in this case to obtain the desired harmonic series. Therefore the multivibrators were compounded, the first operating with a frequency ratio of 5, whose fundamental frequency was therefore 10 kc, and the second with a frequency

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ratio of 10, whose fundamental was consequently 1 kc. Provision was made for a small coupling coil to be inserted in the first multivibrator circuits so that the series of 10-kc harmonics could be utilized external to the assembly. The complete wiring diagram of the assembly is given in Fig. 4, on which the values of the various circuit constants are shown as well as the normal ranges of supply voltages, through which stable control was obtained.

With careful adjustments of the condensers C, *within the control range*, stable operation has been attained for the following ranges of supply voltages, the frequency ratio being 50 for the entire assembly:

e_c from 1.8 to 2.8 volts, at 50 kc
E_f from 3.5 to 6.0 volts
E_p from 90 to 135 volts (or more)

While ranges of supply voltages such as given above have been attained many times, the ranges shown on the diagram are more indicative of nominal conditions, in which no excessive care has been used in making the adjustments. Under these conditions no difficulties are encountered in operation while "trickle charging" the filament and plate batteries from the commercial 60-cycle power line.

The foregoing observations indicate that by injecting a 50-kc voltage into the circuits of the compound multivibrator it is possible to establish a highly stable condition in which the fundamental and all the harmonics of both multivibrator elements are constant in frequency to the same degree that the injected voltage is constant. Moreover, the fundamental output at 1 kc bears every indication of being precisely controlled by the injected voltage. However, the exact mechanism of this control effect has not been analyzed, and we still have no direct experimental evidence that, when the multivibrator operates in a condition which has the physical characteristics of being "controlled," its fiftieth harmonic actually bears a constant phase relation to the control oscillation.

As a final critical test of this point, simultaneous observations of the control oscillation and the fundamental constituent of the multivibrator were made by means of a Braun tube. The circuit arrangement is indicated in Fig. 3A. Two tetrode amplifiers were used, one coupled loosely to the quartz oscillator furnishing

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the control voltage, and the other to the ouptut of multivibrator A or multivibrator B. The output circuits of the amplifiers were carefully tuned so as to emphasize the fundamental frequencies of the standard oscillator and the multivibrator. When the circuits were not accurately resonant, the patterns indicated by the Braun tube were highly distorted. For the operation of the compound multivibrator, with a frequency ratio of 5 in the first and of 8 in the second, the pattern given by the Braun tube appeared as sketched in Fig. 3B.

The form of the pattern clearly indicated the frequency ratio of 40, remaining entirely stationary. As the tuned circuits were changed, the phase of one frequency component could be shifted with respect to the other, of course, but with the circuits

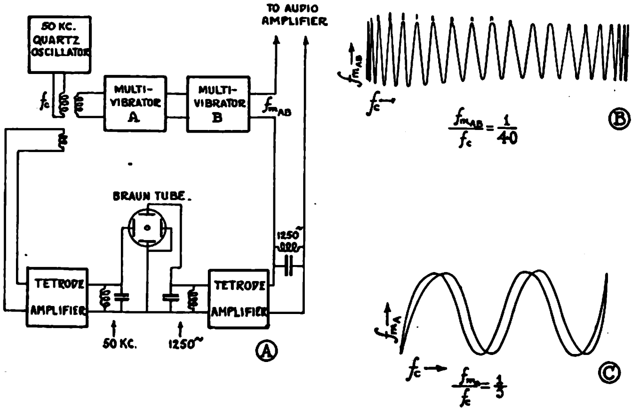

Fig. 3

fixed no visible fluctuations in frequency, phase, or amplitude took place. Upon changing either the resistances or the capacities of the multivibrator circuits, *keeping within the control range*, a change of phase was indicated by the pattern. This change of phase was "permanent,"—that is, varying the multivibrator constants displaced the phase and produced a new form of pattern, which form persisted until some further change was made.

A similar test, made between the frequency of the quartz oscillator and the output of the first multivibrator, gave a pattern which indicated the frequency ratio of 5, with no visible changes in phase or amplitude, indicated by Fig. 3C. As before, changing the multivibrator constants *within the control range* had the effect of changing the phase between the two voltages acting

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on the Braun tube, but for fixed values of the constants the phase remained fixed over indefinite periods. In both of these cases, variation of the circuit constants to a point where the control order changed indicated that a maximum phase variation of not more than 30 deg. (between the control wave and the synchronized harmonic of the multivibrator) could take place before the multivibrator jumped out of control at the desired order. In this test the multivibrator was brought to the edge of a transition range and the amplifiers were then adjusted to obtain the "in phase" pattern on the Braun tube screen. The multivibrator constants were then changed across the control range to the opposite transition point, the resulting change in phase being noted.

When in control, the controlling frequency may be varied by slight amounts, the multivibrator remaining in control. As the variation in frequency is made, the multivibrator suffers a slight phase displacement with respect to the standard frequency, but as long as the phase is not displaced beyond the limit of 30 deg., approximately, the phase takes up the new value and remains constant.

Continuous observations over periods of several hours have not indicated any perturbations of frequency, phase, or amplitude of sufficient duration to register on the Braun tube. After making the amplifier circuit adjustments necessary to obtain a distinctive pattern, on which any slight drift in phase would be readily observable, intermittent observations over periods of days have failed to show any change whatever in the phase of the synchronized multivibrator harmonic with respect to the control voltage.

We conclude from these observations that this state of oscillation of the multivibrator in the presence of a high-frequency injected oscillation which we have called the "controlled" oscillation throughout the discussion, is in fact an oscillatory state wherein the appropriate multivibrator harmonic bears a constant and permanent phase relation to the injected oscillation. Moreover, the frequency of this harmonic and hence of the fundamental follows any drift or perturbation in the frequency of the control oscillation without appreciably altering the phase relation or dropping out of synchronism even for a single cycle. The unsymmetrical multivibrator therefore constitutes a stable

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submultiple generator suitable for converting the vibrations of the piezo oscillator into the rotation of a clock train.

4. The Piezo Oscillator

The circuit of the 50-kc piezo oscillator is shown in Fig. 4. The piezo element *P* is a normal-cut quartz bar, vibrating in the direction of its length; its approximate dimensions are 5.6 x 2.2 x 1.4 cm. Fig. 5 shows the complete assembly of the measuring equipment. On the lower shelf are the batteries and chargers. On the upper, left to right, are the 50-kc temperature-controlled crystal, crystal oscillator, first isolating amplifier, 10-kc multivibrator with coupling coil, second isolating amplifier, 1-kc multivibrator, 1-kc amplifier, and synchronous motor with clock train. The mounting is shown in Fig. 6. The bar is balanced on a metal support *C* having a felt wedge on its upper surface and confined between heavy metal condenser plates *B* by the screw *E*, carrying a felt pad, the whole assembly being mounted on a base of insulating material *D*. The holder and quartz bar are mounted in a constant-temperature chamber *Q*, Fig. 4. This constant-temperature unit consists of an aluminum casting of high heat capacity containing a cavity for the quartz condenser; the casting is surrounded by an insulating layer, which is in turn enclosed in a closed metal box *R*. Heat is supplied from coils distributed around this second metal box, the heating circuit being closed through a bimetallic thermostat positioned, with the heating coils, outside the second container. The whole is enclosed in a heavily insulated outer box of wood, *W*.

The quartz bar is maintained in oscillation by the triode *T*, operated from plate circuit batteries which are independent of the multivibrator batteries.

The quartz oscillator circuit is one of the general class in which sustained oscillations are impossible without the quartz bar. This type of circuit was chosen because experience with numbers of piezo oscillators used as secondary standards has indicated that the frequency of the oscillator is in general more precisely determined by the mechanical vibration of the quartz if no other periodic element is present which can support sustained oscillations independent of the mechanical vibrator. For this reason the regenerative element *G*, which is added to augment the regeneration through the grid-plate capacity of the triode, is a pure resistance. The load in the plate circuit, consisting of

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inductance L shunted by capacity C, is far removed from resonance at the frequency of the quartz bar. It merely represents a convenient method of obtaining the high inductive reactance required to sustain oscillations. The coil has an inductance of 20 mh and the capacity is normally set at about 0.0003 μf, thus providing an inductive reactance of approximately 15,000 ohms. This reactance is still approximately proportional to the frequency in the operating region. The capacity C was added, not to augment the voltage available from the oscillator, since an excess of amplification is provided in order to avoid the necessity of large amplitudes in the oscillator circuit, but merely to provide a means for making fine corrections on the oscillator frequency.

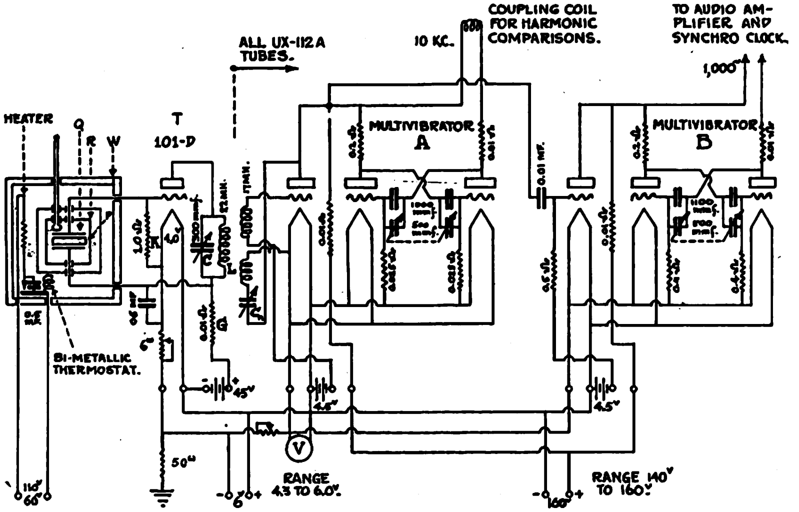

Fig. 4—Schematic Wiring Diagram of Controlled Compound Multivibrator.

The damping coefficient of the quartz bar itself has not been determined, but the characteristics of the whole oscillator in the actual operating region are as given in the following table. The figures here are the fractional change in frequency, positive or negative, for an increase of two per cent in the specified voltage or circuit element.

Two Per Cent Increase in:	Fractional Change in Frequency
Plate Voltage	-0.02×10^{-6}
Filament Voltage	-0.04×10^{-6}
Resistance R_p	$+0.8\times10^{-6}$
Resistance R_g	-0.1×10^{-6}
Capacity	-3.0×10^{-6}

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The temperature coefficient of the quartz bar and holder combined is -5.4×10^{-6} per deg. C.

5. Determination of an Unknown Frequency Between 500 and 1500 kc

The specific purpose of this development was to provide means for accurately adjusting a secondary standard oscillator to one of the even decimal frequencies between 500 and 1500 kc. These secondary standards are piezo oscillators including quartz plates mounted in holders with adjustable air gaps. The plates are adjusted to within one-tenth of one per cent by grinding, and the final setting is made by adjusting the air gap with a screw

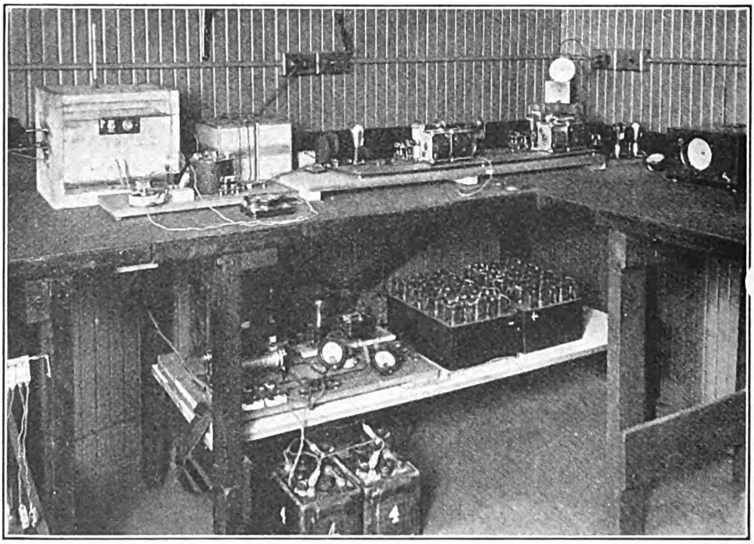

Fig. 5

provided for this purpose which is then locked. The process of referring the adjusted frequency to a standard time interval is carried out by coupling the oscillator under test loosely to the calibration coil of multivibrator *A* (see Fig. 4) which has a fundamental of 10 kc, then coupling a third oscillator and detector loosely to the combination, and listening with telephone receivers to the output of this detector as the oscillator under test is adjusted to beat zero with the appropriate harmonic of the multivibrator. This harmonic can be identified without difficulty with a subsidiary wavemeter, although there is usually not the

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slightest ambiguity as to the order of the harmonic, since the preliminary adjustment of the test oscillator by grinding the quartz is carried out with the aid of a heterodyne wavemeter calibrated to an accuracy of 0.1 per cent and the beat note between the oscillator under test and the multivibrator harmonic is normally well within the audible range before final process is begun. Since a third frequency is superposed on the output of the multivibrator and the test oscillator, the frequency of the test oscillator can be adjusted aurally with a precision defined by about one beat in five seconds between it and the multivibrator.

When this adjustment has been made, the frequency of the oscillator under test is known directly in terms of the frequency of the quartz bar at the time of the measurement, and thus in terms of the standard time interval, to a precision which depends upon how closely the frequency of the quartz is maintained at its average value.

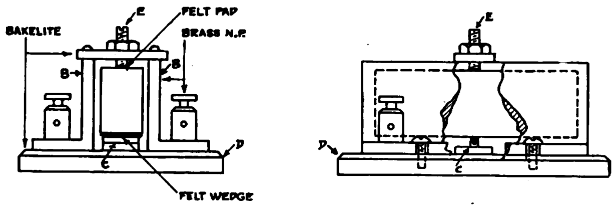

Fig. 6—Quartz Bar Mounting.

The standard time interval now employed in this work is one mean solar day, as indicated by radio time signals from the U. S. Naval Observatory. The oscillation counter consists of a small synchronous motor geared down through a clock train to a large second hand making one rotation per minute. The synchronous impulse motor has a toothed rotor of 120 teeth, and is driven by two U-shaped magnets around which the drive coils are wound. The gearing is such that the second hand indicates true solar time when the driving current has a fundamental frequency of 1 kc per second. The dial reading is compared visually with the standard time signals with a probable error of not over 0.05 second, the mean being derived from ten consecutive observations. Thus the mean frequency over each 24-hour interval is observed with a probable error slightly over one-half part in a million. The use of a longer standard time interval to reduce

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the error in observing the mean frequency will not be justified until substantial reductions are made in the probable short-period perturbations from the mean frequency.

The fluctuation of the frequency during the first ten days in August, 1928, is shown graphically in Fig. 7. Each point in this curve indicates the average frequency during the preceding 24-hour interval except in two cases, where the interval was 48 hours. The horizontal line marked "Mean Frequency 50,007.6 cycles, per second" is the average for ten days. The maximum deviation from the 10-day mean is 0.36 cycle during the fourth and fifth days.

The limitations at present imposed upon the commercial standards to be calibrated require that they be adjusted to an accuracy of ten parts in one hundred thousand or better. The

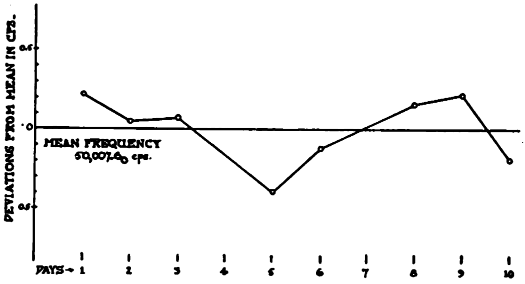

Fig. 7

24-hour mean frequency of the calibration oscillator is known to well within one part in one million. The battery voltages are held within such limits that the short-period fluctuations attributable to them are probably negligible compared with the fluctuations due to temperature. A definite correlation was observed between the temperature and the long-period fluctuations of Fig. 7. It is reasonable to assume that the momentary perturbations from the 24-hour mean are not greater than the maximum fluctuation from the mean observed in a long interval, which is of the order of eight parts in one million. Thus if the calibration is performed with a precision defined by 0.2 cycle in 500,000 or more, the normal accuracy at present attainable in adjusting the oscillator under test is about eight parts in one million which is safely within the current requirements. The piezo oscillator has recently been adjusted to a mean frequency of 50000.15 cycles per second, and in most oscillator calibrations the correction to 50

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kc is neglected, allowing the adjustment to zero beat of the oscillator under test.

The variations in mean frequency of the piezo oscillator as indicated in Fig. 7, while insignificant as regards present requirements, are of course much too large for a precision standard, and are not representative of the possible performance of the system under the most favorable conditions which modern technique of temperature and circuit control may afford.

In an effort to improve the performance of the standard, a new temperature control box was built, though use was still made of the bimetallic thermo-regulator. The fluctuation of the frequency during the ten-day interval from November 4, 1928, is shown graphically in Fig. 8. The improvement over the performance shown by Fig. 6 is apparent, but the results still leave

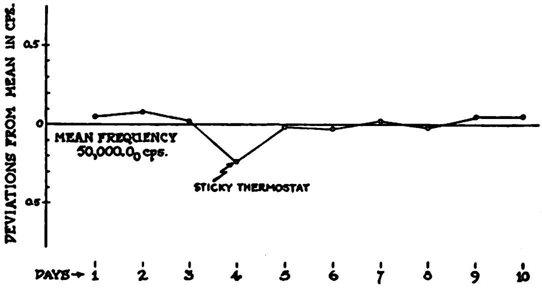

Fig. 8

room for considerable effort. Outstanding lines of development to be pursued are (1) provision of temperature control within one or two hundredths of a degree; (2) replacement of present insulation in the quartz-bar holder by a vitreous material showing negligible distortion with age at constant temperature; (3) improvements in the clock mechanism and in methods of checking against the time signals to provide a higher degree of accuracy in measuring the time interval.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

A SYSTEM FOR FREQUENCY MEASUREMENTS BASED ON A SINGLE FREQUENCY*

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Summary—*The method here described is suitable for calibration of either piezo oscillators or frequency meters in terms of an accurately standardized temperature controlled piezo oscillator. A radio-frequency generator is adjusted and maintained at 10 kc in terms of the standard piezo oscillator. The correct adjustment of this generator is shown by a special form of beat indicator which gives both visible and audible indication of the frequency difference between the piezo oscillator and a harmonic of the 10-kc generating set. A second auxiliary generator may then be set to a harmonic of the 10-kc generating set, such as some frequency assignment in the broadcasting band. The correct setting of the auxiliary generator is also determined by means of the special beat indicator. The piezo oscillator to be tested will usually be found to give an audible note in the telephone receivers connected to it. The audible note represents the difference in frequency between the piezo oscillator and the auxiliary generating set. The frequency difference is determined by comparison with a calibrated audio-frequency generator.*

A SIMPLE method of measuring the frequency of a piezo oscillator is to adjust a radio-frequency generator so that it has the same frequency as the piezo oscillator to be tested, and then measure the frequency of the generator in terms of a calibrated frequency meter. This measurement can be repeated at several harmonics if desired, and the fundamental frequency of the piezo oscillator determined in terms of several coils of the frequency meter. In the method here described, the frequency of the piezo oscillator is determined in terms of an accurately known frequency standard, and the result is not dependent upon the calibration of a frequency meter as an intermediate step in the measurement.

At the time work was begun on the development of this system, in March, 1928, the procedure followed in testing a piezo oscillator for a broadcasting station consisted in determining when the quartz plate was adjusted to the frequency assigned to the station. As the frequency assignments end in

* Dewey decimal classification: R210. Original manuscript received by the Institute, November 19, 1928. Publication approved by the Director of the Bureau of Standards of the U. S. Department of Commerce.

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tens between 550 and 1500 kc, it is apparent that an electrical system giving harmonics spaced 10 kc apart could be used to give the desired frequencies without final measurement upon a frequency meter of extreme precision and accuracy.

A radio-frequency generator for 10 kc can readily be constructed using a vacuum tube. The output of such a generator has a large number of harmonics which may be used in measurements at frequencies many times the fundamental. A generator of this type, however, cannot be relied upon to maintain its frequency adjustment accurately enough for use as a standard, in terms of which standards for broadcasting stations are to be adjusted. A piezo oscillator, however, will maintain its frequency with a high degree of precision if it is maintained at a constant temperature. A temperature-controlled piezo oscillator is therefore a desirable standard to use in maintaining a 10-kc generator accurately upon its frequency.

The system herein described is best explained by reference to Fig. 1 which shows schematically the apparatus as used.

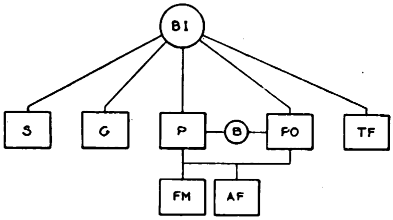

Fig. 1—Schematic Diagram of Apparatus Used in Frequency Measurements. *S*—temperature-controlled piezo oscillator; *G*—10-kc generator; *P*—adjustable frequency power generator; *PO*—piezo oscillator to be tested; *TF*—electrically driven tuning fork; *FM*—frequency meter; *AF*—audio-frequency generator; *BI*—aural and visual beat indicator; *B*—beat indicator.

The system includes: a temperature-controlled piezo oscillator *S*, the frequency of which is accurately known, in terms of which the measurements are made; a 10-kc vacuum-tube generating set *G*, capable of very precise frequency control and giving a very constant frequency; a radio-frequency vacuum-tube generating set *P*, covering the range of frequencies desired; the piezo oscillator to be calibrated, *PO*; an electrically-driven tuning fork of known frequency, *TF*; a calibrated frequency meter, *FM*, for determination of the order of the harmonic of generator *P* in terms of generator *G*; a calibrated audio-frequency generator *AF*, capable of precise adjustment; a special form of com-

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bined beat indicator and generating detector *BI*, by means of which the operator maintains the various elements of the system in the frequency relationship desired; a beat indicator, *B*, which can be used for adjusting the piezo oscillator *PO* precisely to the frequency or harmonic of generator *P*.

In brief, the general method of operating the apparatus is as follows. The frequency of the 10-kc generator *G* is controlled manually by changing the setting of a very small variable condenser and is held at a frequency a harmonic of which produces "zero beat" with the fundamental or a harmonic of the standard piezo oscillator *S*. This condition is determined by means of a beat indicator *BI*. The power generator *P* is then set so that it is at zero beat with a suitable harmonic of the generator *G*, as indicated by beat indicator *BI*. After this adjustment generator *P* is operating at the required frequency. The piezo oscillator to be tested is then adjusted to give zero beat with generator *P* as indicated by beat indicator *B*. The frequency meter *FM* is used only to determine the order of the harmonics or to measure frequency differences. The adjustment of the piezo oscillator is based directly on a standard piezo oscillator.

The low-frequency generator *G* is a Bureau of Standards Type O auxiliary generator described in Letter Circular 187,[1] with the addition of a number of fixed condensers and a small two-plate variable air condenser. This generator employs a type 201-A tube. It was found that by using one of the largest coils and an additional capacity of about 0.012 μf, a frequency of 10 kc could be obtained. After enclosing the generator in a box to eliminate sudden temperature changes due to air currents, it was found to hold its frequency adjustment very satisfactorily. The beat indicator circuits give audible indication of the slightest frequency shift. Three different piezo oscillators with frequencies of 25, 100, and 200 kc have been used at different times as the standard in terms of which the 10-kc generator was adjusted.

In order to emphasize some of the merits of the method, reference is made to Fig. 2, which gives a diagram of the connections for a simple beat indicator where the outputs of the two generators *G* and *H* are led into the indicator circuits having

[1] Specifications for portable auxiliary generator, Bureau of Standards Type O, Bureau of Standards Letter Circular 187.

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coupling coils *C* and *E*, crystal detector *D*, and milliammeter *A*. Whenever generators *G* and *H* have nearly the same frequency or a simple harmonic relation exists between their frequencies, the pointer of the milliammeter *A* will vibrate, indicating the frequency difference or error in the harmonic relationship. Fre-

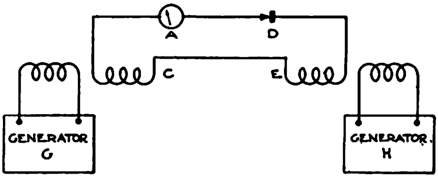

Fig. 2—Circuit for Simple Beat Indicator which Operates When Generator *G* Is Nearly Adjusted to Frequency or A Harmonic of Generator *H*.

quency differences up to a few cycles per second can be observed with the milliammeter of the beat indicator, Fig. 2. A careful adjustment of the frequency of one of the generators will then cause the pointer to vibrate more and more slowly until its deflection does not change. While very precise frequency adjustments can be made in this manner, there may be some tendency for the frequency of the standard to be changed slightly, because of the presence of the coupled indicating circuit. Efforts were made to eliminate the possibility of errors of this type, which lead to the beat indicator shown in Fig. 3.

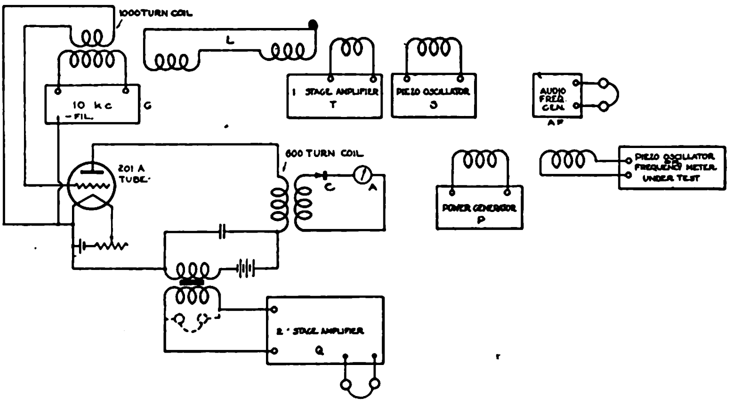

Fig. 3—Diagram of Special Aural and Visual Beat Indicator with Associated Apparatus Used in Frequency Measurements.

This indicator gives visible indications on the milliammeter *A* with very loose coupling between the standard frequency

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generator *S* and the low-frequency generator *G*. When a shielded piezo oscillator is employed at *S*, Fig. 3, a one-stage radio-frequency amplifier and the link circuit *L* are required. The use of this form of coupling does not affect the frequency of the piezo oscillator *S*.

Essentially the beat indicator circuit shown in Fig. 3 is a generating detector circuit. A 1000-turn honeycomb coil in the grid circuit couples to the generator *G* and to the output of the piezo oscillator *S*. A 600-turn coil in the plate circuit couples to a coil of 8 or 10 turns in series with a crystal detector and d.c. microammeter *A*, which gives the visible indication. An audio-frequency transformer is connected in the plate circuit as shown. Telephone receivers or a two-stage amplifier are connected to the secondary winding of the transformer.

To operate the beat indicator the approximate setting of the 10-kc generator is first found by listening in the telephones connected in the plate circuit of the 10-kc generator or as shown by the dotted lines, in which case the two-stage amplifier would be disconnected, or by listening in telephones in the output of the amplifier. After adjusting the 10-kc generator to zero beat, with the telephones connected to the output of the amplifier, a rather faint high note may be heard, or, if the power generator, *P*, Fig. 3, is operating near some harmonic of the 10-kc generator rapid pulsations in the beat note between the two generators are heard. This renders it much easier to adjust the 10-kc generator correctly. By careful adjustment of the tuning control of the 10-kc generator this high note will fluctuate in intensity, and when the fluctuation becomes slow enough the pointer of the milliammeter *A* will be found to vibrate in step with the fluctuation. By further adjustment of the tuning control the pointer can be kept from moving. At this point the 10-kc generator *G* can be shifted by further slight adjustment so that there is either a high-pitched faint note or absolute quiet. The latter condition occurs when the generator is correctly adjusted with respect to the frequency standard used. If the generator gets out of adjustment by as much as one cycle in a minute, the operator will detect it by hearing the high-pitched note appear in the telephones. Slight errors in this setting may not be important unless a high order harmonic is to be used. Errors in adjustment of the order of a few hundredths of a cycle are immediately apparent and the generator can be readjusted.

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The sensitiveness of the visible beat indicator is shown by the fact that when the coils of an unshielded piezo oscillator and 10-kc generator were coaxial and from two to three feet apart, visible beats of the beat indicator milliammeter *A* (Fig. 3) were readily obtained. Type 201-A tubes were used in both the piezo oscillator and the 10-kc generator. The link circuit *L* was not used in this case. In other words, the coupling to the piezo oscillator *S* was extremely loose, which meant that there was no danger of the frequency of the piezo oscillator (which was used as the standard) being changed by the other circuits. Thus the beat indicator system could not affect the frequency of the standard.

Having adjusted the 10-kc generator accurately, the next step is to adjust the third generator, *P*, in Fig. 3,called the power generator, to the desired harmonic of the 10-kc generator. A vacuum-tube detector and amplifier and coil system coupling the two generators would give beat notes in telephones connected in the output circuit of the amplifier, but it has been found unnecessary to add this extra circuit when working up to 1500 kc using a 10-kc fundamental. The telephones in the two-stage amplifier *Q* in Fig. 3 will also serve to indicate the correct adjustment of the power generator. When working at the higher frequencies in the broadcast band, beats which otherwise would not be heard are made audible by placing a pick-up coil and the phone leads in the vicinity of the power generator.

The frequencies of the power generator ending in tens of kilocycles give a zero beat which cannot be set accurately by aural methods because of the uncertainty of the setting for true zero beat. The frequencies ending in 2 1/2, 5, and 7 1/2 can be set accurately by matching beat notes in the telephone receivers, which are beats of beats.

Although the 10-kc points cannot be set precisely, they can be accurately determined by matching a beat note in the phones with a corresponding note from a tuning fork. If the frequency of the fork is known, it is only necessary to make the measurement on one side of zero beat. If this method is adopted, great care is necessary to insure that the desired fork frequency is used, as the measurement may be made by matching the generator beat note with a submultiple or harmonic of the tuning fork. In order to guard against this error it has been found convenient to make an approximate zero beat measurement first.

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Then a measurement is made with the tuning fork by detuning the generator so that the frequency heard in the telephones due to the beat note between the 10-kc generator and the generator is the same as that of the tuning fork. It is necessary that the frequency meter used have sufficient resolving power to show an easily read difference in the condenser settings for these two frequencies, and the curves should be large enough to allow a correct determination of what harmonic of the fork is used.

If it is desired to prepare an enlarged frequency meter curve, it is evident from the above paragraph that a large number of points can be obtained for the calibration over as small a frequency range as 10 kc. Assuming that a 400-cycle electrically-driven tuning fork is available, in the range from 200 to 210 kc, for example, the following points could be very accurately determined using the beat notes 199.200, 199.600, 200.400, 200.800, 202.500, 204.600, 205.000, 205.400, 207.500, 209.200, 209.600, 210.400, and 210.800. These thirteen points permit the drawing of a large accurate curve.

The circuits described above have been used for calibration purposes between 40 kc and 1500 kc, and beat notes have been obtained up to 2500 kc, spaced 10 kc apart. In the lower part of the frequency range beat notes were obtained 2500 cycles apart when using a 25-kc piezo oscillator and a 10-kc fundamental from generator *G*.

The measurement of the frequency of a piezo oscillator can be made by the method described with the addition of an accurate audio-frequency generator. For example, suppose the frequency of a piezo oscillator is to be accurately measured, and it is known that its frequency is approximately 600 kc. The power generator *P* of Fig. 3 is set to the sixtieth harmonic of the 10-kc generator as determined by a common type of frequency meter. Then when listening in the phones of the piezo oscillator a beat note will, in general, be heard, which is the difference between the frequency of the piezo oscillator and the generator. This beat note is matched by a similar note from the audio-frequency generator, from the calibration of which the audio frequency is determined. Whether this audio-frequency is to be added to or subtracted from 600 kc is determined by listening in the piezo oscillator telephones and noting whether the frequency of the power generator must be increased or decreased to pro-

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duce zero beat. Check measurements can be made at other harmonics or subharmonics of the piezo-oscillator frequency.

It has been found desirable for checking purposes to take a reading on a frequency meter of high resolving power at the time that the audio-frequency beat note measurement is made. A certain frequency value is obtained from the calibration of the frequency meter. Generator *P* is then adjusted to zero beat with the piezo oscillator under test, and another reading taken on the frequency meter. These two frequency meter readings give a frequency difference which serves as a check on the frequency measured with the audio-frequency generator, and also indicates whether the frequency as determined by the latter generator is to be added to or subtracted from the frequency as given by generator *P*, which, as previously stated, is based on the standard piezo oscillator.

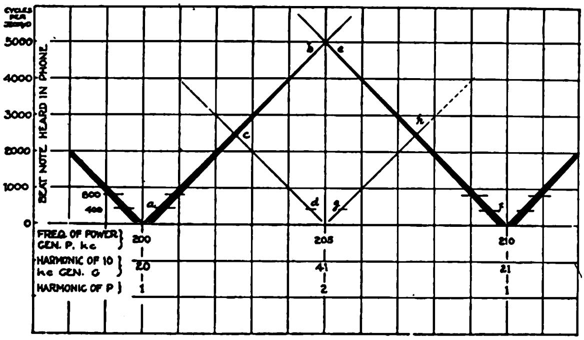

Fig. 4—Chart Showing Example of Relations between Beat Notes Heard in the Special Beat Indicator Telephones, with the Harmonics of Generators *G* and *P* Indicated.

The frequency of a quartz plate mounted in an adjustable mounting can readily be altered to the desired value by setting the power generator to the correct harmonic and then adjusting the quartz plate mounting to give zero beat. A beat indicator can be brought into use for a more accurate adjustment than is possible by use of the telephones alone.

Fig. 4 illustrates how the various beat notes heard in the telephones can be interpreted, i.e., whether the frequency to which the power generator is set ends in 10 or in some other number. The diagonal lines represent the beat notes heard in the phones, and the breadth of the line its relative strength. Thus it is seen that the beat notes occurring near the frequencies

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ending in 10 are very loud or strong. Zero beat being indefinite, a setting is taken by matching with the note from an electrically-driven tuning fork as described above. Increasing the frequency of the power generator from 200 kc, for example, as indicated in Fig. 4, the beat note gradually gets higher and fainter, line *ab*, until at a frequency of the order of 204.5 kc the beat note is quite high, but a lower note, line *cd*, will also be heard which becomes lower as the other becomes higher. Soon after the low note has become silent the high note will be heard to pulsate, and, with critical adjustment of the generator, beats with another beat note will be heard which permit a very accurate setting of the generator. This setting gives one of the 5-kc points and is readily distinguished over a range of several hundred kilocycles by the aural method described. If the frequency of the power generator is increased further the other high-frequency beat note, line *ef*, becomes lower and stronger, and a weaker beat note, line *gh*, becomes higher and fainter. Further change in the power generator brings in a loud beat note and finally silence or zero beat in the telephones. This is another 10-kc point. The beat note represented at *b* is 5000 cycles when a 10-kc fundamental is used. When changing the frequency of the power generator from *a* to *b*, if conditions are right, another beat note may be heard at *c* and a precise setting made at this point which will be midway between the 10-kc point and the 5-kc point. This setting is more difficult to obtain than the 5-kc point because it is indicated by the beats between two 2500-cycle notes of much different intensities. When using a piezo oscillator having a fundamental frequency of 25 kc, the 2500-cycle notes are readily distinguishable. Other beat notes may be heard and can be set accurately but may not be readily usable. Points 5 kc apart are much closer than are usually required.

An explanation of the various beat notes heard in the phones can be given by further reference to Fig. 4 where harmonic numbers are indicated for the three zero-beat frequencies given. The 200-kc frequency or fundamental of the power generator beats with the twentieth harmonic of the 10-kc generator, and the 210-kc fundamental of the power generator beats with the 21st harmonic of the 10-kc generator. When the frequency used is 205 kc, the second harmonic of the power generator, or 410 kc, beats with the 41st harmonic of the 10-kc generator.

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As previously explained, the 202.5-kc and 207.5-kc points are obtained by listening to the beats between the beat notes heard in the phones. Beat notes other than those indicated in Fig. 4 may be heard in the phones, but those described are usually more than are required for calibration purposes.

While this system or method was intended chiefly for calibration of piezo oscillators, it has been found to be ideal for frequency meter calibration work. The system has the following advantages:

(1) Extreme accuracy and precision obtainable.
(2) Large number of calibration points possible.
(3) Flexibility and ease of operation.
(4) Ease of working up final results from experimental data.

With reference to (1), the accuracy with which the frequency of the standard is known and maintained will determine the accuracy with which the measured frequencies can be determined, because of the extreme precision possible in working with beat notes. This precision is far in excess of even specially built frequency meters. While the 10-kc generator is not held automatically to its frequency, yet the characteristics of the circuit are such that the operator is immediately aware of any change in its frequency and can readjust it.

Item (2) above is important because the system permits calibration of piezo oscillators for broadcasting stations for any frequency in the broadcasting band. It extends the usefulness of a standard piezo oscillator since it allows the exact setting of many harmonics which could not be obtained from the piezo oscillator alone. This fact is of great importance in the calibration of special frequency meters capable of extreme precision and having high resolving power such as one used in the radio laboratory of the Bureau of Standards. The large number of calibration points possible can be seen from the example cited above. This fact also adapts the circuits to the measurement of the frequency of any piezo oscillator by final measurement of an audio frequency by means of an audio-frequency generator.

With reference to (3), after the apparatus is correctly adjusted it is very easy to operate and adapts itself to any frequency calibration. The frequencies to be adjusted can be spaced uniformly as desired by setting what has been called the 10-kc generator to frequencies other than 10 kc, such as 5 kc, 15 kc,

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20 kc, 25 kc, etc. For example, it may be that points spaced 100 kc apart are desired. The 10-kc generator in that case would be set to 100 kc.

With reference to (4), data on frequency meter calibration can be worked up and results obtained almost by inspection if the frequency of the piezo oscillator used as a standard is such as to permit the setting of the 10-kc generator accurately on 10 kc, and provided the experimental data is carefully taken and the same procedure is used in matching the fork frequency. Ten-kc and 5-kc points are readily distinguished experimentally, and existing curves can be checked rapidly or corrected by adding 10 kc to successive 10-kc points starting from a known point. By adding 10 kc to successive 5-kc points starting at a known 5-kc point these frequencies are rapidly determined.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

RECEIVING SETS FOR AIRCRAFT BEACON AND TELEPHONY*

By

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Summary—*Special light-weight radio receiving sets of high sensitivity have been designed adapted for the reception on aircraft of signals from radio beacons and ground radio telephone stations. A statement of the important considerations involved in the aircraft reception of signals from Government-operated beacons and telephone stations along the civil airways in this country, such as frequency range, tuning means, volume control, vacuum tubes, amplifier characteristics, size, weight, and performance specifications, is given.*

Design details for three receiving sets of slightly different types, with numerous characteristic and performance curves, are discussed, with a brief discussion of the results of practical flight tests.

INTRODUCTION

THE conditions surrounding the use of radio receiving equipment installed on aircraft impose many special considerations of both design and performance. The specifications for aircraft sets, therefore, differ materially from those employed for other purposes.

Such sets at this time are of three types. One type fulfills military requirements where a rather broad band of frequencies must be covered while in flight. Another type is required to receive the signals from beacons and ground radiotelephone stations which the United States Government will supply along the civil airways under the air commerce act. The third type is adapted to the high-frequency (above 1,500 kc) field. Most of the activities of the past have, in this country, been confined to sets of the first type mentioned. The other two fields are new.

European radio developments applied to commercial aviation have been under way for several years. Resulting practice has centered around a narrow band of frequencies between 315 and 350 kc for aircraft services. Radio beacons, particularly those

* Dewey decimal classification: R521. Reprinted from Bureau of Standards Journal of Research, October, 1928. Publication approved by the Director of the Bureau of Standards of the U. S. Dept. of Commerce.

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erected in the United States for aiding marine navigation, have adopted adjacent frequencies near 300 kc. The 1927 International Radio Conference recognized these established practices and the International Convention allocated the bands 285 to 315 kc for all radio beacon services and 315 to 350 kc for aircraft communication services, 333 kc being adopted as an international calling and distress frequency for aircraft. The directive radio beacons and weather broadcasting services planned for our civil airways will, therefore, be confined to these two bands.

These bands being adjacent and covering a total range of only 65 kc permit of simplifications in receiving set specifications for this type of service. As part of its research and development work[1] on directive radio beacons and radiotelephony for aircraft under the air commerce act, the Bureau of Standards has developed such receiving sets, the design elements of which are treated in this paper.

Requirements for Beacon and Telephone Receiving Set

The dominating requirement in design is that of small weight and physical dimensions. This refers not only to the set, but also to the batteries requisite for its operation. Every design element must be considered in the light of this requirement.

The set must be capable of installation anywhere on the aircraft with simple remote controls when placed at an inaccessible point. The tuning system should, therefore, be of a uni-control type over the entire 285 to 350 kc band for the sake of convenience.

As an aircraft is in very rapid motion, the distance over which signals are traveling is constantly changing. A simple and rugged volume control having a uniform action from zero to maximum signal strength and capable of remote installation from the receiving set is consequently necessary.

Sufficient shielding of the set must be provided to limit interference from the engine ignition system to that induced in the antenna structure.

Rugged construction to withstand the continuous mechanical vibration on aircraft must be provided. Vacuum tubes must be used that do not introduce microphonic noises, which, if present, effectively increase the already prevailing noise level.

[1] J. H. Dellinger and H. Pratt, "Development of Radio Aids to Air Navigation", Proc. I. R. E., **16**, 890–920; July, 1928.

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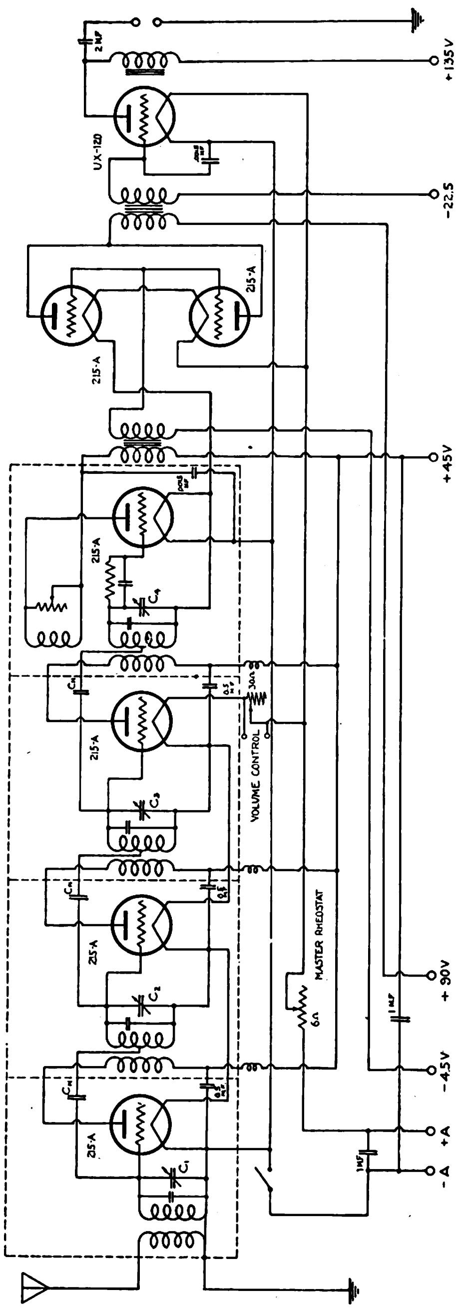

Fig. 1—Circuit Diagram of Completely Shielded Receiving Set.

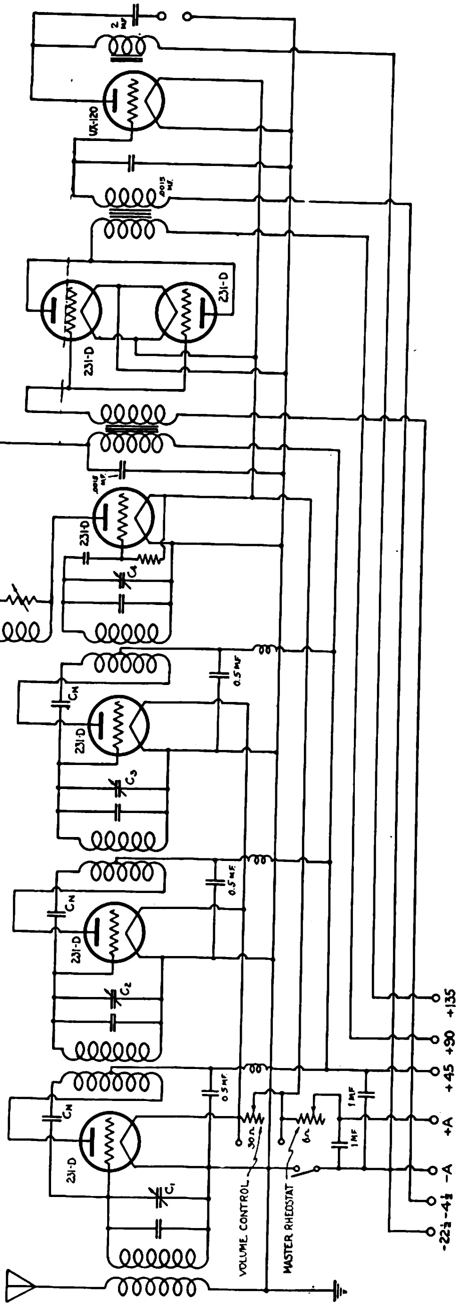

Fig. 2—Circuit Diagram of Partially Shielded Receiving Set Using 3-volt Tubes.

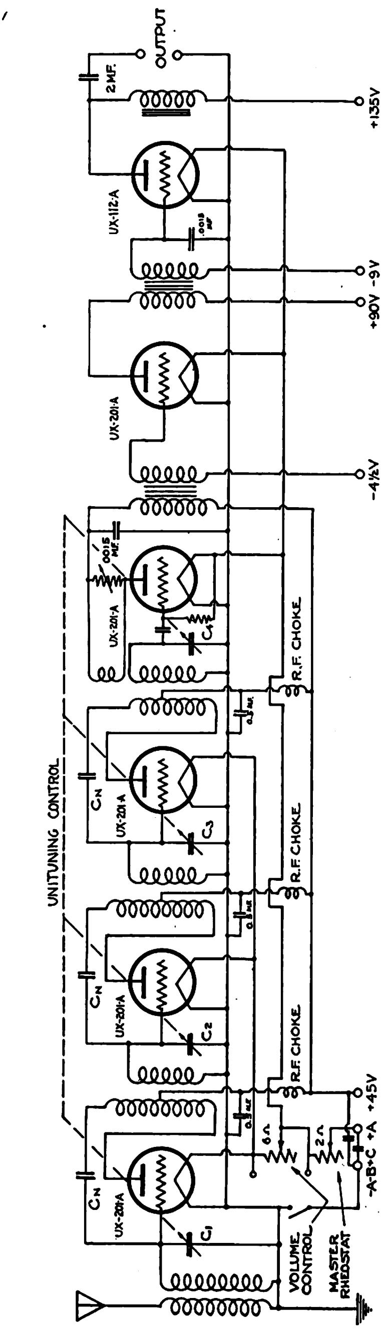

Fig. 3—Circuit Diagram of Partially Shielded Receiving Set Using 5-volt Tubes.

The set must be capable of operating a beacon course indicating instrument. Its output must efficiently supply a maximum of 10 volts of audio-frequency power into a load impedance of from 4,000 to 7,000 ohms, to provide an ample margin of signal strength. The direct-current plate supply must be isolated from the output circuit to prevent the observer from receiving shocks when wearing headphones, and to prevent the setting up of a polarization in the course indicator's magnetic system.

The audio amplifiers and output circuit must provide a nearly uniform amplification over a frequency range of 40 to 3,000 cycles. The beacon signals are modulated in the range 40 to 120 cycles, and a voice frequency range from 200 to 3,000 is needed for good intelligibility.

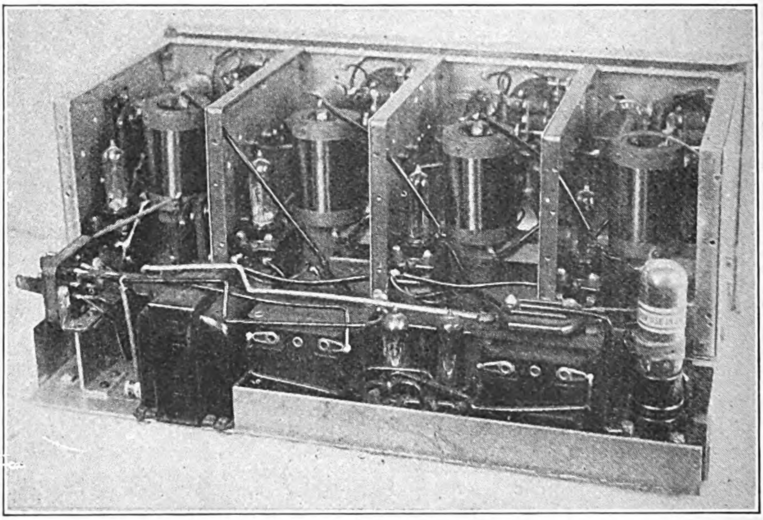

Fig. 4—Interior of Completely Shielded Receiving Set.

Both the sensitivity and selectivity must be high. High sensitivity is desirable to permit short vertical pole antennas to be employed because they eliminate direction errors and remove the danger of weighted trailing wires. Good selectivity is necessary to enable close spacing of frequency channels.

The choice of vacuum tubes has an important bearing on the total weight and bulk of the equipment. Three types of receiving sets will be described; two of which employ 1-volt and 3-volt tubes, respectively, where the total A battery current is of the

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order of 0.5 ampere; the third using 5-volt tubes where the *A* battery must supply 1.5 amperes. By employing the lower voltage types both the *A* storage battery can be reduced in size and the advantages of vacuum tubes having small physical dimensions secured. The standard low-voltage tubes offered on the market, however, are not sufficiently sturdy for aircraft use, being often mechanically weak and usually causing microphonic noises. Special tubes having these disadvantages to a less degree were employed, but owing to their special nature the third set

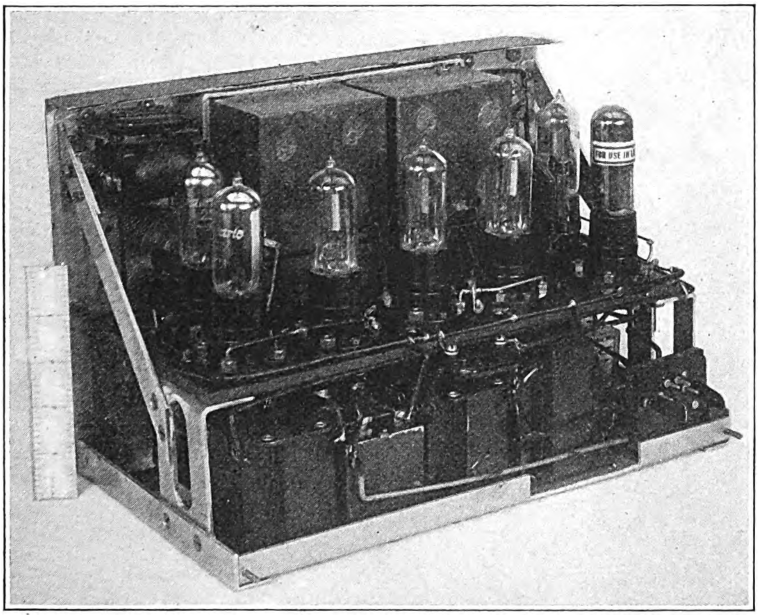

Fig. 5—Interior of Partially Shielded Receiving Set Using 3-volt Tubes.

equipped with the standard 5-volt tubes was also designed. There is much need for better tubes for aircraft sets, and it is hoped that commercial designs will be improved.

Receiving Set Design

Figs. 1, 2, and 3 show the circuit diagrams of the three receiving sets designed to meet the above requirements, and Figs. 4, 5, and 6 are corresponding photographs. Their physical characteristics are tabulated below.

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TABLE I

WEIGHTS AND DIMENSIONS OF RECEIVING SETS WITH ASSOCIATED BATTERIES

Receiving Set No.	Weight of Set (pounds)	Dimensions (inches)	Weight of batteries (pounds)
1	18	$9\frac{1}{2}$ by $7\frac{1}{4}$ by $16\frac{1}{2}$	19
2	16	$8\frac{1}{2}$ by 9 by 14	19
3	17	$6\frac{1}{4}$ by 8 by 13	25

Sensitivity, selectivity, and unicontrol features are met by employing three or four tuned radio-frequency amplifier circuits with gang variable air condensers. Some regeneration in the detector circuit is provided. After many trials superheterodyne sets were eliminated from consideration owing to the presence of two tuning

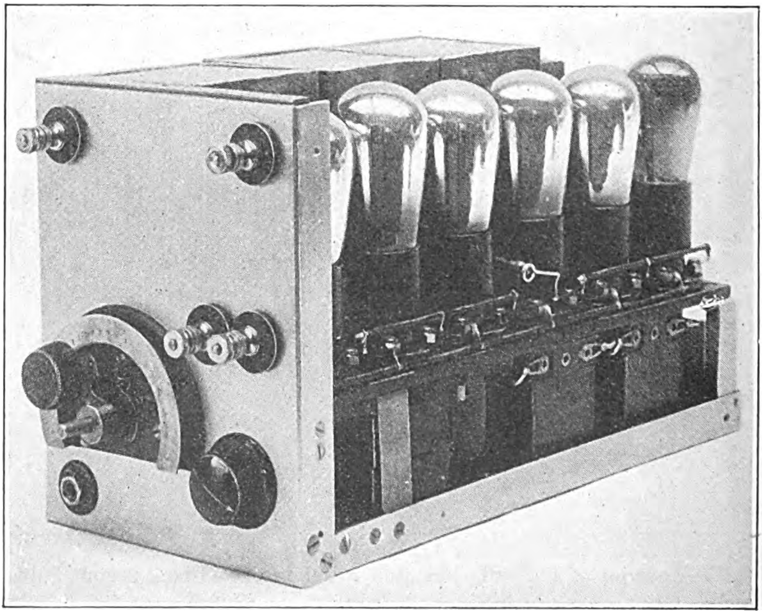

Fig. 6—Interior of Partially Shielded Receiving Set Using 5-volt Tubes.

positions for each frequency. A few hours of flight would usually bring the aircraft within close range of stations having frequencies such as to create interference by beating with the intermediate frequency or its harmonics. A pilot operating a radio-equipped airplane in bad weather, when the radio services are needed, must devote all his attention to piloting, and the radio set must convey the information it receives without attention and with-

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out interference from other signals or it will lose usefulness and possibly become a hazard.

1. Design of Radio-Frequency Transformer (Interstage). Radio-frequency transformer theory has for some time been reduced to mathematical treatment.[2] A brief consideration of the analysis involved is necessary to a clear understanding of the actual design procedure employed. Fig. 7 shows the commonly accepted equivalent circuit corresponding to a stage of radio-frequency amplification employing tuned transformer coupling between stages; R_1 represents the tube plate resistance plus the radio-frequency resistance of the transformer primary, L_1 the primary inductance, M the mutual inductance between windings, and R_2, L_2, and C_2 the constants of the secondary circuit.

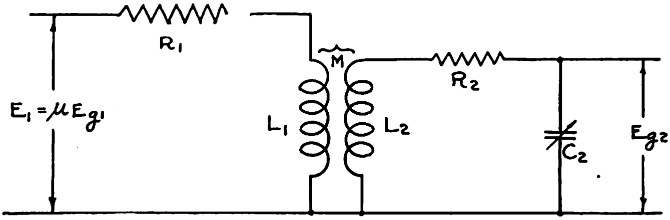

Fig. 7—Accepted Approximate Equivalent Diagram of a Stage of Radio-Frequency Amplification.

In transformer design those values of the three-circuit parameters X_1, X_2, and M which will make the voltage gain per stage a maximum at the desired frequency are of interest. If X_2 alone is varied (by means of the tuning condenser C_2) the voltage gain is a maximum when

$$\frac{X_2}{X_1}=\frac{\omega^2M^2}{Z_1^2} \tag{1}$$

If, now, either X_1 or M is varied, X_2 being simultaneously varied in order to satisfy the condition expressed by (1) for each value of X_1 (or M), the voltage gain will reach an optimum value when

$$\frac{R_2}{R_1}=\frac{\omega^2M^2}{Z_1^2} \tag{2}$$

Equations (1) and (2) may be combined into the simpler relationship

$$\omega^2M^2=Z_2Z_1 \tag{3}$$

[2] Victor G. Smith, "A Mathematical Study of Radio-Frequency Amplification," Proc. I. R. E., 15, 525–536; June, 1927. G. W. Pierce, "Electric Waves and Oscillations", McGraw Hill Book Co. (Inc.).

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When this relationship is fulfilled the voltage gain K is given by (4)

$$K_{\text{optimum}} = \frac{1}{2} \frac{\mu}{\sqrt{R_1}} \frac{\omega L_2}{\sqrt{R_2}} \tag{4}$$

and is the maximum amplification that can be obtained with a given tube and coil.

The actual transformer design closely followed the above theoretical considerations.

Two secondary coils (single-layer solenoids) were chosen, having suitable values for $\frac{\omega L_2}{\sqrt{R_2}}\cdot$ The first coil consisted of 135 turns of No. 28 single silk enameled wire wound on a 2¾-in. diameter coil (winding length 2¾ in.) and having a self-inductance of 1 millihenry, a radio-frequency resistance of 18 ohms at 290 kc, and a value for $\frac{\omega L_2}{\sqrt{R_2}}$ at the same frequency equal to 430. The second coil was of 250 turns of No. 34 enameled wire wound on a 1¾-in. diameter coil (winding length 1¾ in.) and having a self-inductance of 2 millihenries, a radio-frequency resistance of 70 ohms at 290 kc, and a value for $\frac{\omega L_2}{\sqrt{R_2}}$ equal to 435 at this frequency. It should be noted that to cover the necessary frequency range of 285 to 350 kc with a comfortable margin on each side a tuning condenser of 0.00035 μf maximum capacity is required with the first coil, and a condenser of 0.00015 μf maximum capacity with the second coil.

Since $\frac{\omega L_2}{\sqrt{R_2}}$ is very nearly the same for each coil, the same optimum voltage amplification should be obtained using either coil in conjunction with a given tube. The second coil, however, has the considerable advantage of compactness. Furthermore, since its selectivity is less than that of the first coil (due to its greater resistance), it becomes somewhat more adaptable for use in a unicontrolled multistaged amplifier.

To determine the optimum value of mutual inductance between windings for each transformer suitable primary forms were chosen and test amplifiers constructed comprising the particular transformer under test together with a given tube. A known volt-

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age of the desired frequency was applied to the input terminals of the test amplifier and the transformer secondary circuit tuned by condenser C_2 for maximum output voltage (as indicated by a tube voltmeter); these measurements being taken for a series of values of mutual inductance. For each measurement, therefore, (1) above is satisfied; while at the particular value of M which yields the optimum output voltage, (2) is, in addition, fulfilled. This is the point of optimum mutual inductance.

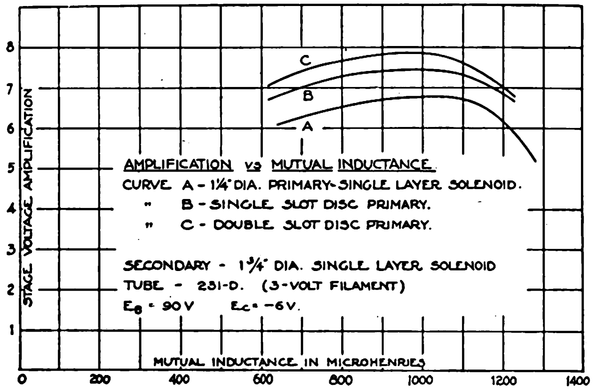

Fig. 8—Amplification vs. Mutual Inductance, 1¼-in. Diameter Primary Single-Layer Solenoid.

Curves A of Figs. 8 and 9 show the variation of voltage amplification with mutual inductance for the 1¾ and 2¾ in. diameter

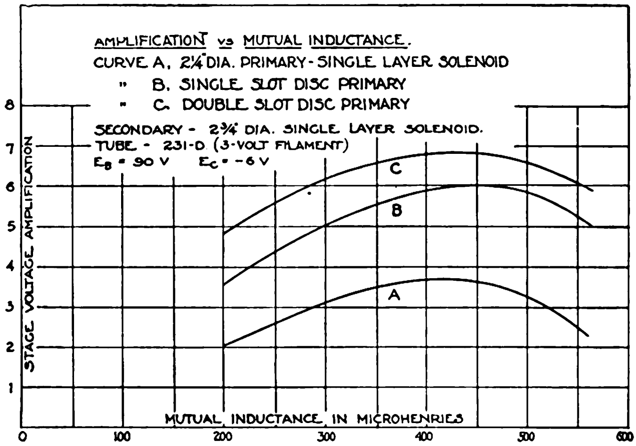

Fig. 9—Amplification vs. Mutual Inductance, 2¼-in. Diameter Primary Single-Layer Solenoid.

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transformers, respectively. The amplification obtained with the two transformers is seen to be decidedly different despite the nearly equal values of $\frac{\omega L_2}{\sqrt{R_2}}$ in each case, and the exactly similar conditions of test; the superiority of the smaller transformer residing in the relatively lower capacitive coupling existing between its windings by virtue of their smaller physical dimensions. That this is true may be seen from a consideration of curves *B* and *C* of Figs. 8 and 9. Curves *B* were obtained with the primaries wound in single slotted disks inserted at the low potential ends of the secondaries, thereby reducing the effective capacitive couplings. The resultant increase in amplification is greater for the 2¾-in. diameter transformer because of the greater reduction in its capacitive coupling. Distributing the primary windings

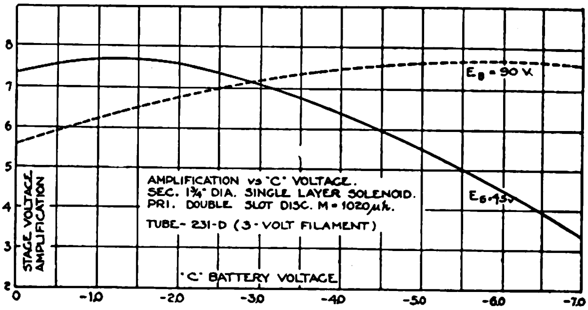

Fig. 10—Amplification vs. *C* Voltage.

in two slots in the disk form causes a still further improvement in amplification, the percentage increase being again in proportion to the decrease in capacitive coupling between windings.

The curves of Figs. 8 and 9 were taken for 90 volts on the plate with a grid bias of -6 volts. It was decided to reduce the plate voltage sufficiently to preclude the use of a *C* battery and thereby to remove the need of filtering in the grid return leads. Fig. 10 shows that the same amplification as in Figs. 8 and 9 can be obtained using 45 volts on the plate and $-1\frac{1}{2}$ volts on the grid. A further reduction of the *C* voltage to zero results in but a small loss in amplification.

Fig. 11 gives the amplification-frequency characteristics for the test transformers; the optimum mutual inductance as found from the curves of Figs. 8 and 9 being provided in each case. Curve *A* is for the 2¾-in. diameter transformer with double-

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slotted disk primary; curve *B* for the 1¾-in. diameter transformer with single layer primary; and curve *C* for the 1¾-in. diameter transformer with double-slotted disk primary. All the important

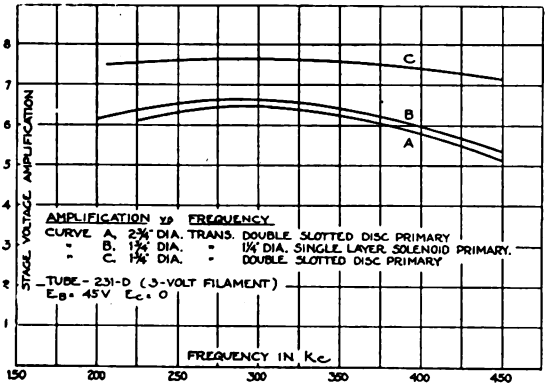

Fig. 11—Amplification vs. Frequency.

constants of these transformers are collected in Table II. A study of the amplification-frequency characteristics shows that they are highly satisfactory over the desired frequency range (285 to 350 kc).

TABLE II
TRANSFORMER CONSTANTS

Transformer No.	Primary Winding	Outside Diameter	Primary Turns	Size of Wire	Secondary Winding	Outside Diameter	Secondary Turns
		Inches				*Inches*	
1	Double-slot disk..	2⅝	135	No. 38 S. C. C.	Single-Layer solenoid	2¾	135
2	Single-Layer solenoid	1¼	250	No. 34 enamel.	..... "	1¾	250
3	Double-slot disk..	1½	250	No. 38 S. C. C.	..... "	1¾	250

TABLE II (cont'd.)
TRANSFORMER CONSTANTS

Transformer No.	Size of Wire	L_1 (μh)	L_2 (μh)	R_2 at 290 kc	M (μh)	Capacitive Coupling between Winding $\mu\mu$f	$\omega^2 M^2$ at 290 kc	$Z_1 Z_2$ at 290 kc, approximately
1	No. 28 S. C. C. enamel	2,930	950	18	450	18	0.67×10^6	0.34×10^6
2	No. 34 enamel..	1,100	1,990	70	1,065	18	3.77×10^6	1.27×10^6
3	....do.........	4,290	1,990	70	1,020	6	3.44×10^6	1.37×10^6

In Fig. 12 are shown the resonance curves of the above three transformers, used in conjunction with a 3-volt tube. Curves *A*,

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B, and *C* correspond to the transformers of curves *A*, *B*, and *C*, respectively, of Fig. 11. It is interesting to note the mutual inductance effect upon the selectivity. Fig. 13 indicates how the selectivity varies for four values of *M*, using the 1¾-in. diameter transformer with double-slotted disk primary.

2. Effect of Capacitive Coupling. It was shown above that the presence of capacitive coupling between windings adversely affects the efficient operation of a transformer. When capacitive coupling exists, the approximate equivalent circuit diagram be-

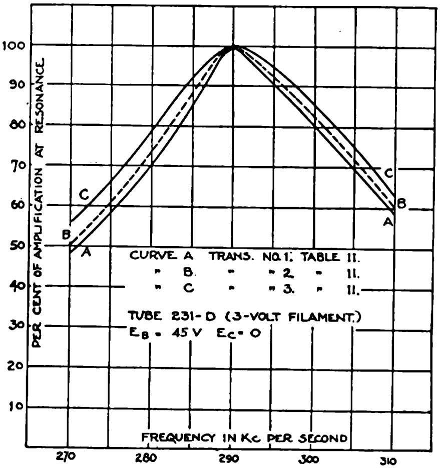

Fig. 12—Amplification vs. Frequency.

comes as shown in Fig. 14 rather than that of Fig. 7. The conditions for optimum voltage gain and the actual magnitude of this optimum gain as found from a solution of this circuit are also different.

A close agreement between experimental results and the theoretical equations based on the circuit of Fig. 7 cannot, therefore, be expected unless the effect of capacitive coupling is minimized. As computed from (3) the optimum mutual inductance for transformer Nos. 1, 2, and 3 of Table II should be 315, 615, and 615 μh, respectively. Actually the values obtained in the laboratory are 450, 1,065, and 1,020 μh, respectively.

The effect of capacitive coupling between transformer windings is thus not only to reduce the magnitude of the optimum voltage gain but also to require considerably more mutual inductive coupling in order to obtain the best operating condition.

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3. Method of Neutralizing. The system of tube grid-plate capacity neutralization adopted depends considerably upon the type of transformer employed. The neutralizing schemes in common use are essentially bridge circuits in which two coils, the grid-plate tube capacity, and a neutralizing condenser constitute the four bridge arms. The bridge maintains a balanced adjustment for all operating frequencies, provided the two coil arms are tightly coupled. When the coupling between the two coils is small a leakage inductance is introduced into one of the bridge arms, thus making the balance dependent upon the frequency.

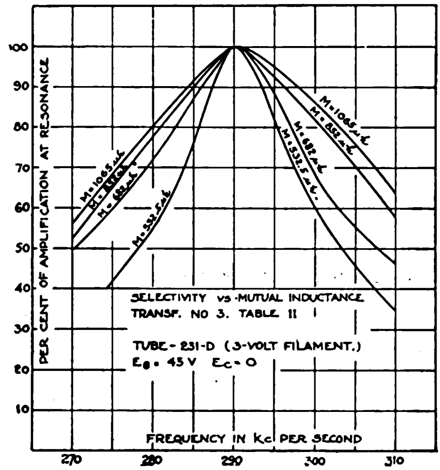

Fig. 13—Selectivity vs. Mutual Inductance.

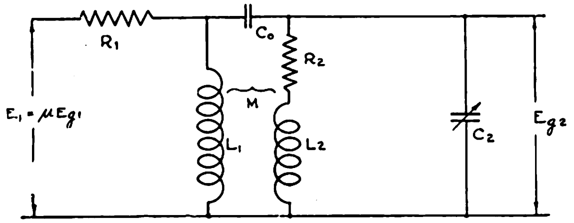

Fig. 14—Closer Approximate Equivalent Diagram of a Stage of Radio-Frequency Amplification.

With several neutralizing systems the primary and secondary windings of the transformer may be used as the two coil arms, provided there is sufficient coupling between them. In the case of the 1¾-in. diameter transformer with single-layer primary,

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a coupling coefficient of the order of 70 per cent obtains. The system employed can, therefore, be that indicated in the circuit diagram of Fig. 1.

In the case of the 1¾-in. diameter transformer with disk primary (used in the receiving set of Fig. 2), however, the coupling coefficient between windings was about 35 per cent. The use of an auxiliary coil coupled either to the primary or secondary winding is therefore necessary. A simple solution consisted in placing the auxiliary coil in a third slot in the primary disk form thus obtaining nearly unity coupling to the primary. An alternative method, due to Miller, is also applicable in this case. In this method the four bridge arms are made up of the tube grid-plate capacity, a neutralizing condenser, and two auxiliary condensers, one of which may be the tube grid-filament capacity. The closeness of coupling between primary and secondary windings is here obviously immaterial.

4. Design of Antenna Transformer. No mention has as yet been made concerning the design of the antenna coupling trans-

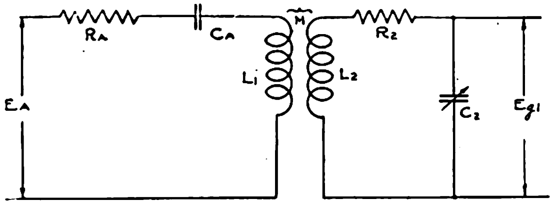

Fig. 15—Approximate Equivalent Diagram of Antenna Tuning System.

former. The theory and procedure of design is essentially the same as for interstage transformers. The equivalent circuit is that of Fig. 15, where C_A is the antenna capacity and R_A the equivalent antenna resistance. Receiving set sensitivity of such order was desired as to make possible the use of a rigid antenna of about 10 feet in length, with the airplane itself as a counterpoise. The capacity of such a system is of the order of 25 to 50 $\mu\mu$f. Using these values, the same transformer as used for interstage coupling was found highly satisfactory.

5. Design of Regenerative Coil. To obtain the order of amplification desired, it was decided to use three stages of radio-frequency amplification with regeneration on the detector. The circuit arrangement of Fig. 16 was used in the design of the regenerative coil. A known modulated voltage E_1 (of the desired radio-frequency) was applied to the input circuit and the voltage

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E_2 measured first without regeneration and then with regeneration. The ratio of the second to the first value of E_2 is the amplification due to regeneration. These measurements were taken for a series of values of regenerative coil turns, adjusting the shunting resistance R to just below oscillation at each point (where necessary). For the particular transformer considered, it was found that the best value of turns was approximately 40 per cent of the number of turns in the secondary winding.

6. Shielding and Arrangement of Parts. Both complete and partial shielding methods were employed in the tuned amplifier stages. Reference to Figs. 4, 5, and 6 will indicate the shielding arrangements. Completely shielding the stages as in Fig. 4 was found to be difficult owing to the involved mechanical assembly. Aluminum was selected to reduce weight, but resulted in objectionable eddy-current losses. Partial shielding using copper

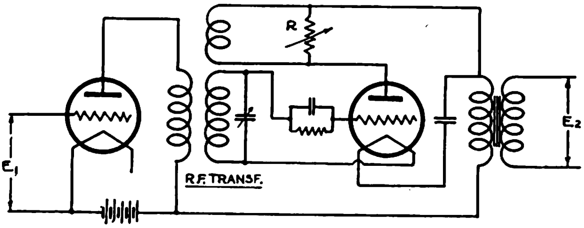

Fig. 16—Test Circuit Arrangement for Measuring Amplification Due to Regeneration.

containers housing the radio-frequency transformers, with proper care taken to make all high potential leads as short as possible, resulted satisfactorily. This latter method was applied as shown in Figs. 5 and 6.

The copper containers maintain an effective clearance of ⅞ in. around the coil windings, and are of square cross section. The coil inductances are decreased by about 20 per cent without any material increase in resistance.

To confine the radio-frequency currents as far as possible to the shielded portions of the set, battery leads were filtered where necessary. The selection of a low-plate voltage served not only to reduce the demand on the B battery, but to permit the radio-frequency stage grid returns to be connected directly to the negative filament terminals, eliminating a common C battery impedance. All other precautions in arrangements and wiring used in good design practice were observed. A multiple conductor

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battery cable, equipped with a detachable plug connector, was provided to allow a rapid change of sets to be made when in practical operation.

A small separate panel containing a combination *A* and *B* battery voltmeter, filament switch, head telephone jack, and volume control was provided for mounting near the pilot, preferably on the instrument board. The volume control consists of a rheostat connected in the radio-frequency amplifier tube filament supply circuit.

The unicontrol tuning feature makes it possible for the operator to tune to any station frequency within the two bands. In this case an antenna coupling arrangement is needed which makes the setting of the antenna tuning condenser at a given frequency independent of the antenna length. This is taken care of by an antenna trimming condenser.

Adjustable stops to allow the unicontrol handle to be rapidly turned from one predetermined frequency in the 285 to 315 kc band to one in the 315 to 350 kc band were provided.

An alternative tuning arrangement was incorporated in one of the receiving sets built, applicable for the case where only one beacon and one telephone frequency would be encountered on any one airplane run. Two sets of condensers connected in parallel were used in each tuned amplifier stage. Normally with both condensers in circuit the tuning adjustment is made for the beacon station frequency. When a switch is closed by the operator four small automatic relays respond, removing one condenser from each circuit, leaving the set tuned to the telephone station frequency which is in the higher band.

7. Detector. In general, grid rectification has been found considerably more sensitive than plate detection, and was, therefore, used. Laboratory measurements gave 3 megohms as the optimum value for the grid leak and 0.00035 μf as the best value of blocking condenser at the frequencies used. These values are not at all critical.

Sensitivity, however, is not always the deciding factor in the choice of the type of detection to be employed. When used in an airplane not excellently shielded, the ignition interference picked up by the antenna structure may be sufficient to block the detector, even though the incoming signal is small. By using plate detection a considerably greater interference would be necessary for blocking the detector. Since the present form

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of course indicator suggested for the beacon signals is practically immune to interference, it will operate satisfactorily so long as the tubes employed are not blocked. It should be noted, however, that plate detection precludes the use of a transformer in the detector output due to the large tube plate resistance with this form of detection.

8. Audio Amplifier. The low amplification factor of the 3-volt tubes used in the first two models (Figs. 1 and 2) necessitated the use of transformer coupling to provide a degree of audio amplification sufficient to supply the desired voltage to the power tube without overloading the detector. Due to the relatively high plate resistance of these tubes, however, the low-frequency pitches are not amplified in the same magnitude as those of higher frequencies. This is indicated in Curve *A* of Fig. 17

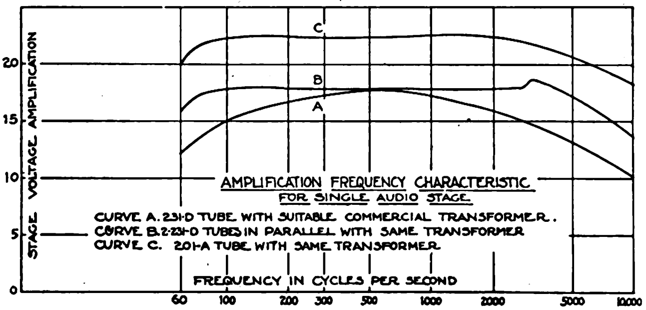

Fig. 17—Amplification-Frequency Characteristic for Single Audio Stage.

which represents the amplification-frequency characteristics for a 3-volt tube used together with a commercial transformer. A large number of commercial transformers were tested, the one finally adopted being chosen for its lightness as well as its electrical characteristics. A much more uniform gain over the desired frequency range is obtained by using two tubes in parallel in the first audio stage in order to reduce the net plate resistance. The amplification-frequency characteristic for this condition is given by curve *B* of Fig. 14. This arrangement was, therefore, used in both the 3-volt receiving sets. Curve *C* shows that when 5-volt tubes are used a satisfactory characteristic is obtained with one tube in the first audio stage.

Using 5-volt tubes a resistance-coupled audio amplifier immediately suggests itself. The third model (Fig. 3) employing 5-volt tubes uses transformer coupling, however, chiefly for the

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sake of flexibility. In this set it is feasible to use either 3 or 5 volt tubes throughout, requiring only a change in the size of the filament control rheostats. The operator can, therefore, decide as to whether the cost of tubes or weight of battery is the chief consideration.

Quite fortunately, the impedance of the visual course indicator is suitable for use with any of the power tubes available. The choice of the power tubes is therefore controlled by the filament voltage and plate battery consumption. A filter arrangement consisting of a choke and condenser is used in the output circuit for reasons previously mentioned. In the case of a resistance-coupled amplifier this filter circuit would also tend to prevent low-frequency oscillation.

9. Over-All Performance. The performance of a receiving set may be indicated by means of three graphs; the first showing the variation of sensitivity with carrier frequency; the second showing the degree of its selectivity at several carrier frequencies;

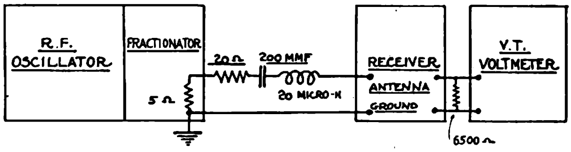

Fig. 18—Test Circuit Arrangement for Measuring Receiving Set Performance.

and the third giving a picture of its fidelity or response to audio frequencies which modulate the carrier to which it is tuned. The conditions under which these tests should be taken have been worked out by the Standardization Committee of the Institute of Radio Engineers for broadcast sets. Those conditions apply here also in large measure, but some departures were found desirable in order to approximate actual use more closely.

The sensitivity curve (Fig. 19) was obtained with the circuit arrangement of Fig. 18. For convenience, 100 per cent modulation at an audio-frequency of 60 cycles was employed. This more nearly approximates actual conditions as low modulating frequencies of the order of 60 cycles are used in the beacon, and the modulation is constant and as near 100 per cent as can be obtained. The curve shows the relationship between the radio field intensity in microvolts per meter necessary to obtain 6 volts

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across the 6,500-ohm load and the carrier frequency to which the receiving set is tuned.

The value of 6 volts was chosen, since this is the maximum voltage needed across the present type of course indicator.

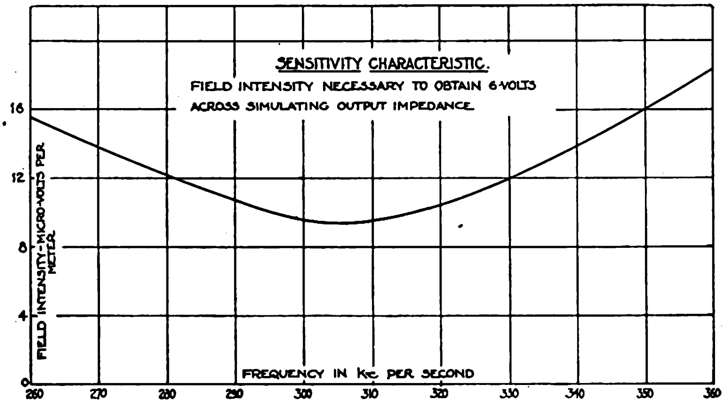

Fig. 19—Sensitivity Characteristic.

Fig. 20 shows the selectivity curves for the set when tuned to 290 and 330 kc. These are, respectively, the approximate beacon and phone frequencies. These curves represent the field intensity in microvolts per meter to give 6 volts across the simulating impedance referred to the carrier frequency.

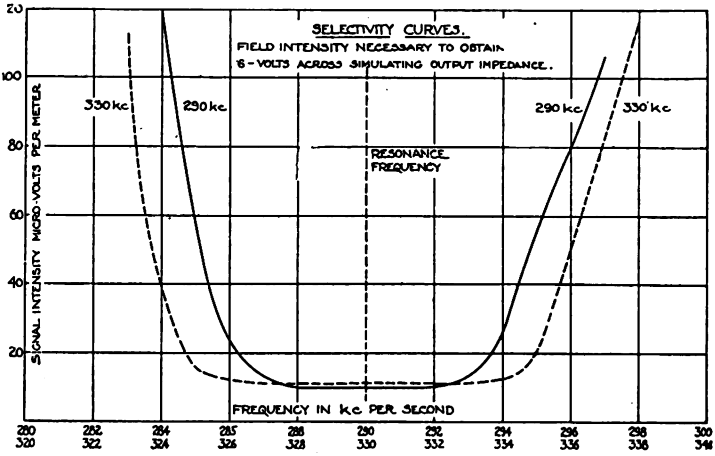

Fig. 20—Selectivity Curves.

Fig. 21 shows the fidelity curve. This was obtained using the audio portion of the receiving set only, and is plotted as the ratio of voltage across the simulating output impedance to that between the grid and filament of the detector tube against audio-frequency.

Twenty volts across the simulating output impedance was obtained before tube overloading occurred.

It should be noted that the curves of Figs. 19, 20, and 21 were obtained with one of the 3-volt sets. When 5-volt tubes are used, measurements indicate that the performance of the radio-frequency amplifier is but slightly altered. The voltage gain in the audio amplifier will, however, be considerably increased due to the greater tube amplification factors.

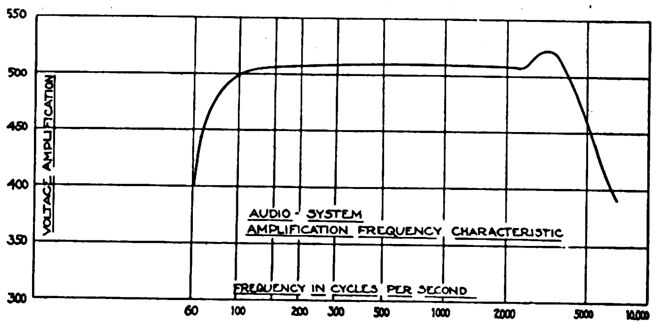

Fig. 21—Audio System Amplification-Frequency Characteristic.

Conclusion

These sets were developed concurrently with the double modulation type of directive radio beacon, the reed type of visual indicator for aircraft, and the airplane vertical pole antenna, all of which are parts of the bureau's research program on aids to air navigation. The development of these receiving sets was found essential because of the special receiving requirements on airplanes peculiar to these air navigation aids. Without receiving sets of high sensitivity, the elimination of the dangerous trailing wire antenna, and the reduction of night beacon course shift errors obtained with the short vertical antenna, would not be possible.

The second set described was extensively used in flight tests for a period of six months. Satisfactory daytime beacon signals were received on an airplane at a distance of 100 miles with an

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antenna 10 feet high, the engine ignition system being adequately shielded.

The writers wish to acknowledge the assistance of L. L. Hughes and D. O. Lybrand in constructing these sets, including several preliminary models.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

AN AIRCRAFT RADIO RECEIVER FOR USE WITH RIGID ANTENNA*

By

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Summary—*An outline is given of the physical and electrical requirements of an aircraft radio receiver suitable for the reception on a rigid antenna of radio beacons and weather service. The design of a special unicontrol receiver calculated to fulfill these requirements is described. Quantitative performance data are presented, with particular attention to the problem of detector overloading when operating a visual indicator from a beacon of the Bureau of Standards type. The corroboration of these data by practical tests is briefly discussed. The paper is concluded with quantitative discussion of the problem of ignition shielding on a particular type of airplane motor.*

THE rapid growth in air transportation of freight, mail, and passengers, together with the development of radio aids to air navigation in the forms of beacon and weather service, has created an urgent need for suitable aircraft radio receiving equipment to operate over a limited band of frequencies. In the 1927 International Convention the 285–315-kc band was allocated for radio beacon services, and the 315–350-kc band for aircraft communication. It is the purpose of this article to describe the physical and electrical characteristics of an aircraft radio receiver designed for the above frequency ranges. This receiver is of practical commercial design, and has successfully withstood the severe conditions of regular operation on mail planes.

The design of the receiver was undertaken over a year ago at the request of the Bureau of Standards. The primary requirement of such a receiver was that it should be suitable for the reception of both visual and aural beacons with a consistent range of 150 miles on signals from a 2-kw ground transmitter, using a rigid pole as antenna not more than two meters in height and the bonded metal members of the airplane structure as a counterpoise. The design of a receiver to meet this unusual sensitivity requirement was undertaken as a progressive step in that program of aircraft receiver development so ably initiated in the work described elsewhere by Messrs. Pratt and Diamond.[1]

* Dewey decimal classification: R521. Original manuscript received by the Institute, January 2, 1929.

[1] This issue of the Proceedings, page 283.

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The severe conditions of vibration and shock, and the great diversity of climatic conditions under which an airplane receiver is used, present certain novel problems of design. Above all else, an airplane receiver must be rugged, light, and compact. The receiver as a whole must be adequately shock-proofed from its mounting to prevent destruction of tubes and wiring by vibration; supplemental shock-proofing of the vacuum tubes within the receiver is also advantageous since the tube elements have natural frequencies of mechanical vibration which may be excited at certain engine speeds. Lightness and compactness in the power supply are highly desirable. A further important requirement of present airplane receivers is that they be capable of installation wherever a vacant spot may be discovered in the plane. For simplicity of tuning and remote-controlling the receiver should have few controls and these should be of ample dimensions and shaped so that a pilot may easily operate them even though wearing heavy gloves. Further, it is highly important that all controls be equipped with some form of locking device to prevent creeping caused by vibration. The problem of rendering radio receiver performance sensibly independent of moisture is acute in designing a receiver for airplane use. A plane flying through fog encounters extremes of humidity seldom met indoors on the ground. The combination of receiver and associated antenna should have sufficient sensitivity to reach the average atmospheric noise level, assuming, of course, that ignition interference from the motor has been reduced to insignificance by shielding. A brief discussion of ignition shielding will be given later in this paper.

It was formerly considered necessary to employ a trailing wire as a receiving antenna. Mechanical objections to this arrangement are obvious, particularly in the case of mail or freight airplanes where the pilot must handle the antenna. Experience with reception of beacon signals has disclosed the further disadvantage of a trailing wire receiving antenna which results from the interception by the antenna of any horizontal components of electric force in the incoming wave. Two types of error are associated with these horizontal components, the "airplane effect" and the "night effect." "Airplane effect"[2] is produced by the inclination of

[2] By the Sommerfeld reciprocity theorem this situation is equivalent to a direction-finding loop on the ground and transmission from a partly horizontal antenna. The error so produced has been recognized for many years. See Ballantine: "Yearbook of Wireless Telegraphy," (London, 1921).

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the antenna when the airplane flies at any angle greater than zero with respect to the beacon course. It can occur irrespective of the presence of a downcoming or reflected wave, and thus may be troublesome in the daytime. It would be entirely eliminated by a wholly vertical antenna system. "Night effect" is produced by the downward wave reflected from the Heaviside-Kennelly layer and may be modified by the ground wave. In general it is increased by increasing length of the transmission path. It manifests itself, in the case of directional transmission to a non-directional receiver, as an apparent wandering of the transmitted beam. When the directional transmitter is a loop system, "night effect" cannot be eliminated by the use of a vertical antenna, although it may be somewhat reduced thereby.

To summarize, the use of a rigid vertical antenna is justified by the elimination of physical hazards and burdens of maintenance on the pilot, as well as by a substantial reduction in the normal beacon errors. Accordingly, such an antenna was recommended by Radio Frequency Laboratories at the outset and the receiving equipment was designed with this in view. Subsequent experience has tended to justify this line of approach.

Owing to constantly changing conditions in the allocated frequency band it is difficult to specify just what minimum degree of selectivity the aircraft receiver should possess. At the present time, the problem of congestion in the 285–350-kc band is not acute, and for satisfactory discrimination the aircraft receiver requires somewhat less selectivity than is required in a modern broadcast receiver. Experience indicates that the order of selectivity corresponding roughly to three tuned circuits in cascade, none of which has a power factor greater than one per cent, is sufficient for present requirements.

A severe demand on fidelity of the aircraft receiver is imposed by the visual type of beacon. In the visual beacon the radio-frequency excitations of the two crossed coils are modulated respectively with two different audio frequencies instead of with dots and dashes as in the aural type beacon. These two audio frequencies differ by about 20 cycles and are usually both between 50 and 120 cycles per second. The aircraft receiver should therefore respond uniformly to modulation frequencies down to 50 cycles per second. For radio-telephonic reception of weather reports the aircraft receiver should maintain substantial fidelity up

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to frequencies of 3000 cycles per second and cut off rather sharply above this frequency in order to reduce noise.

Fig. 1 is from a photograph of a mail plane equipped with the RFL receiver and vertical pole antenna. In this installation the receiver was mounted in the front of the pilot's cockpit so that a remote tuning control was not required. The dash control panel (which will be described later) contains the volume control and is mounted on the instrument board. The front face of the receiver, dash control panel, and reed indicator are visible in Fig. 2, which is from a photograph taken looking forward into the pilot's cockpit. The antenna is a 7/8-in. diameter duralumin tube triangularly guyed to the upper wing and fuselage. In later installations the

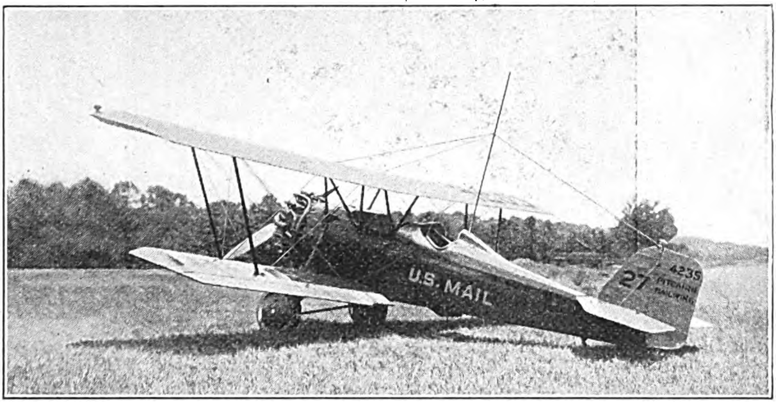

Fig. 1—Installation on Pitcairn PA-5 Mail Plane Showing Short Rigid Vertical Antenna.

tube has been stream-lined to reduce wind resistance. The height of the tube above the fuselage is about two meters. With most planes, this will permit entrance to hangars having a clearance of 14 feet, so that the antenna need never be taken down.

Fig. 3 shows the receiver in its shock-proof mounting, the dash control panel, and a set of reeds for use on the visual indicating type of beacon. These reeds are held in a shock-proof mounting and are accurately tuned with temperature compensation to the two modulation frequencies of the beacon transmitter. They vibrate with equal amplitudes when the airplane is on its course. If the plane strays to the left of its course, the left reed develops a greater amplitude of vibration than the right, and vice versa. The dash control panel includes a volume control,

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which varies the filament current of the radio-frequency amplifier tubes, a double-range voltmeter for indicating the condition of plate and filament batteries, a switch for turning the receiver filaments "on" and "off," two cables, one making connection with the receiver and the other with the power supply, and a jack for

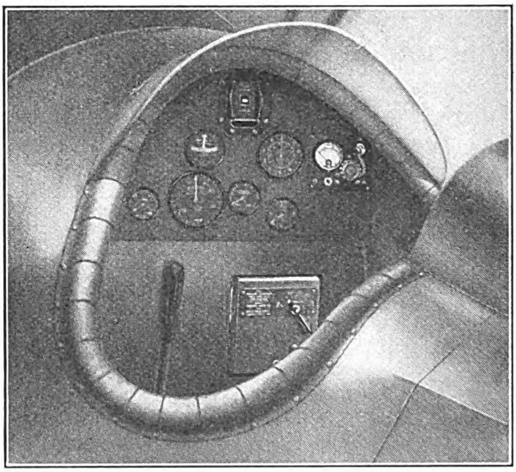

Fig. 2—View of Control Cockpit in Pitcairn Plane and Receiver Installation; Reed Indicators on Instrument Panel, upper center; Receiver Control Panel at right; Receiver Below.

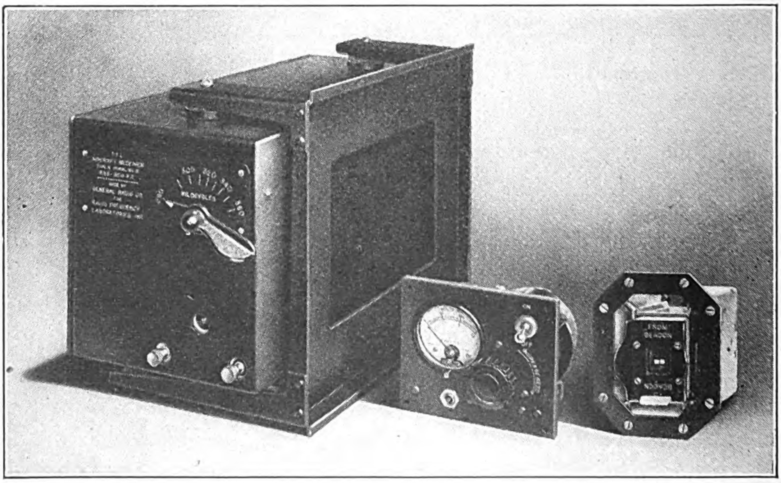

Fig. 3—Model B Unicontrol Aircraft Receiver, Dash Control Panel and Reed Course-Indicator.

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the connection of telephones or reeds. The cable from the control panel to the receiver makes connection with the receiver through a multiple point plug. The mounting frame which supports the receiver is constructed to slide and lock in a track which is permanently fixed in the plane. These arrangements simplify the removal of the receiver from the plane. The front face of the receiver contains antenna and ground binding posts and the single tuning control, about which a direct calibration in kilocycles is engraved. The tuning control is the only control physically associated with the receiver itself, and is therefore the only control for which remote operation need ever be specially provided.

Fig. 4—Special Light-Weight Combined *B* and *C* Dry-Cell Battery for Use with Aircraft Receiver.

A special combined *B* and *C* battery has been developed for use with this receiver. One of these batteries is shown in Fig. 4. This battery is designed to give a life of 100 operating hours at a current drain of 10 milliamperes, which is the maximum drain of the receiver. Dimensions and weight of this battery, together with the dimensions and weights of the various units of the receiving equipment are as follows:

Dimensions

Overall dimensions, receiver in mounting	15 in.×9 in.×6 in.
Dash control panel	4½ in.×4 in.×3½ in.
Reed box in holder	4½ in.×4 in.×6 in.
Combination *B* and *C* battery	10 in.×7¾ in.×3 in.

Weights

Receiver with tubes and shock-proof frame	11 lbs.	0 oz.
Dash contol panel and cables	1 "	14 "
Reeds in shock-proof mounting	1 "	9 "
Combination *B* and *C* battery	10 "	6 "
Antenna pole and fittings	5 "	0 "
Total	29 lbs.	13 oz.

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In almost all instances the airplane carries a storage battery for navigating or landing lights. This battery may be used for the filament supply of the radio receiver. Otherwise, a small 6-volt lead storage battery is provided. Such a battery, especially designed for airplane use, is now made by a number of manufacturers and weighs about ten pounds.

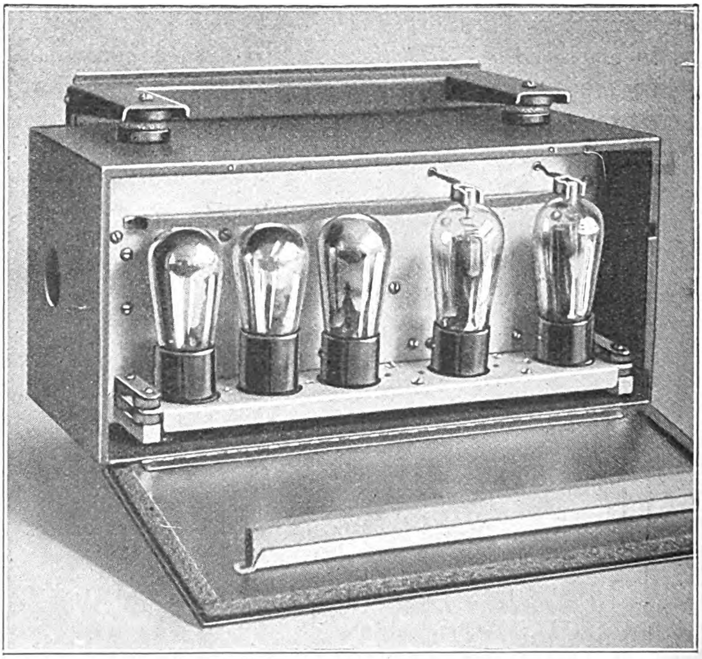

Fig. 5—Tube Compartment of Model B Receiver, Showing Shock Insulation of Tube Girder and Auxiliary Shock-Proof Support at Top of Tubes.

The left side of the receiver is hinged to permit access to the tube compartment. In Fig. 5 this side has been opened to show the arrangement and method of shock-proofing the tubes. The tubes are all mounted on a rigid aluminum channel, which is insulated from the receiver chassis by sponge rubber washers. Two horizontal sponge rubber strips, one fitted to the hinged side and the other to the receiver chassis, bear lightly against the tops of the tubes above the points of maximum diameter of the bulbs, and prevent the tubes from creeping out of their sockets or moving laterally in case appreciable vibration reaches them.

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Felt strips along the edges of the hinged side serve to protect the chassis from dust and moisture.

Five tubes are employed as follows: two shielded tetrodes as radio-frequency amplifiers; two high-μ triodes as detector and first audio amplifier; and a power triode as the second audio amplifier.

There are three tuned circuits operated by a single control. In place of the conventional gang condenser a gang variometer is used. This gang variometer is a rigid unitary assemblage of three approximately cubical cells, each of which houses a variometer. The design of the radio stage is simple and merely involves determining for a chosen size of shield the variometer of required inductance range which has the lowest power factor. Two general types of construction have been successfully employed for the variometer gang: first, a welded construction using pure aluminum sheet; and second, a cast construction using an aluminum alloy coated with copper by the Schoop spray. Fixed condensers molded in bakelite are used in conjunction with the variometers in the tuned circuits.

Tuning by variable inductance is not desirable where a wide frequency range must be covered, but it possesses marked advantages when a frequency band defined by limits of two to one or less is required. It results in a very obvious saving of weight and space. Further, when used in conjuction with an antenna short in comparison with the wavelength, it permits exact single control with no trimming adjustments, when the antenna capacity is made equal to the fixed capacities employed in the other tuned circuits. Identical variometers may then be employed in the radio stages and antenna input circuit.

The use of tetrodes in the radio amplifier results in a radio gain of approximately 50 per stage instead of the gain of 10 or 15 per stage ordinarily obtainable with triodes. A high radio-frequency gain per stage is a more important consideration in an aircraft receiver than in ground sets where lightness and compactness are not vital. For the first audio amplifier, or at least for the detector, the tetrode is not so acceptable because of microphonic noises inherent in the structure of tubes now commercially available.

A detector capable of withstanding considerable overloading is provided in the receiver by the use of plate rectification with

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automatic grid bias.[3] This feature is particularly important when using the receiver in conjunction with the visual-type of beacon indicator. When closely approaching the beacon a pilot may neglect to keep his reed amplitude down to the proper level by adjustment of the volume control; if this condition persists the detector may be so overloaded as to cause the reed amplitude to pass through a maximum and then fall to normal levels. Under

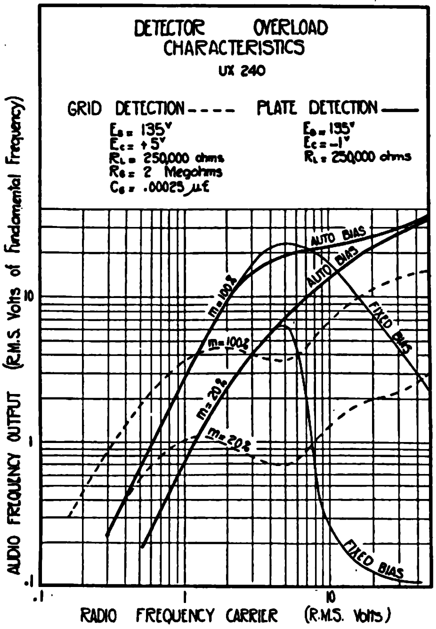

Fig. 6—Comparison of Ordinary Grid and Plate Rectifying Detectors and Special Detector with Automatic Grid Bias Used in Model B Receiver; Fundamental Component of Audio Output Voltage for Modulations of 20 and 100 per cent Against Carrier Voltage.

such conditions the indicated course may be reversed, i.e., the reed of lesser amplitude will indicate the side to which the airplane is off course.

Fig. 6 depicts the overload characteristic of three detector arrangements for two degrees of modulation (m). The curves show audio output of *fundamental* modulation frequency produced by a wide range of radio inputs. Examination of the curves

[3] This arrangement was devised by Stuart Ballantine.

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shows that a somewhat higher output may be obtained with plate rectification than with grid rectification before overload. With either type, however, reversal or lack of course indications may result if the input voltage is increased through values corresponding to the maxima of the curves into the regions of negative slope. This matter is of considerable importance as in the practical operation of the receiver without automatic bias near a beacon, many instances of detector overloading have been observed. The method of preventing overloading employed in this receiver has been found quite satisfactory since it meets the essential requirement for correct indication of course, that the output curve shall always have a positive slope. While the audio amplifier may "overload" before the detector (plate rectification), the necessity for avoiding a negative slope in the detector response is in no way affected thereby, since the amplifier "overloading" at these voltages merely involves a falling-off in response which does not attain a negative slope.

Plate detection with automatic grid bias has other advantages. It is commonly supposed that grid rectification results in greater sensitivity than plate rectification. As a matter of experimental fact the microvolt sensitivity of this tetrode receiver is twice as great with plate detection as with grid detection, although the small-signal detection factor for the latter is about three times as great as for the former. Plate detection exceeds grid rectification in this case because it leaves unaffected the radio gain of the preceding stage, whereas due to electronic conductance grid detection reduces this gain by a factor greater than two to one. For the same reason the selectivity is considerably greater with plate rectification than with grid rectification. The higher output impedance resulting from plate detection does not impair the uniform transmission of low modulation frequencies essential for the visual beacon provided the coupling between detector and audio amplifier is properly designed.

A resistance coupled audio ampl'fier is employed in the receiver. In addition to mechanical advantages of compactness and light weight, such an amplifier has electrical characteristics desirable in the airplane receiver. Among these are low *B* battery drain and uniform transmission of the required modulation frequencies.

Fig. 7 gives the overall performance characteristics of the receiver. Sensitivity is expressed as the 400-cycle, 30 per cent

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modulated microvolt input required in the standard antenna to produce 15 rms volts audio-frequency output across a 4000-ohm resistive load. The standard antenna here considered has 100 $\mu\mu$f capacity and negligible resistance. These constants are characteristic of the rigid antenna for which the receiver was designed. If the effective height of the rod antenna is one meter the receiver will yield an output of 15 volts for field intensities ranging from 5 to 8 microvolts per meter.

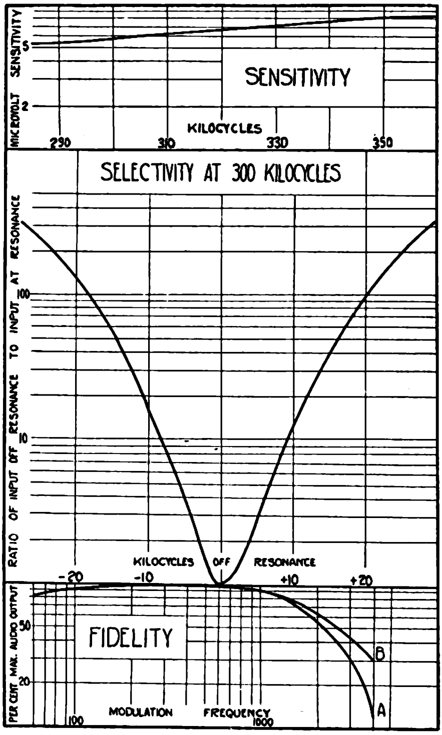

Fig. 7—Performance of Model B Receiver as Shown by Standard Measurements of Sensitivity, Selectivity at 300 kc, and Fidelity (A, of entire receiver; B, with radio-frequency amplifier out).

Selectivity is expressed as the ratio of the radio input voltage off resonance to the radio input voltage at resonance required for the normal audio output. It is important that this ratio be plotted over a range of several frequency channels on either side of resonance in order to present a sufficient basis for judging the discrimination of the receiver. The "skirts" of the selectivity curve are as important as the region immediately about resonance.

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Fidelity is expressed as the percentage of the audio output at any modulation frequency to that at a reference frequency of 400 cycles per second, and is obtained with fixed radio input by measuring the audio output at various modulation frequencies. The overall fidelity thus measured includes all frequency distortions produced in normal operation of the receiver. Curve *B* represents fidelity in the absence of sideband attenuation in the radio-frequency amplifier. It is obvious from a comparison of this curve with the actual fidelity curve (curve *A*) that a major reduction in high audio-frequency response occurs in the radio amplifier. Fidelity and selectivity were measured at a carrier frequency of 300 kc per second; the results are not substantially different at other carrier frequencies in the range.

In addition to the standard laboratory measurements of performance represented by the sensitivity, selectivity, and fidelity curves, extensive tests of the receiver have been made under a wide variety of service conditions with the rigid antenna.

Tests were made early in 1928 with the receiver installed in an airplane belonging to the Bureau of Standards. With the receiver operating at less than full sensitivity, aural beacon signals from Bellefonte, Penna., were received at College Park, Md., a distance of about 140 miles. An installation on a Pitcairn Mailwing affords a satisfactory indication of the range on visual beacon service by producing at Hadley Field, N. J., normal amplitudes of the reed indicators from signals transmitted from College Park, Md., when the College Park transmitter is supplying a current of only 5 amperes to each loop. The two loops are triangular in shape, 70 feet high and 300 feet along the base. Repeated tests on an airplane belonging to the Radio Frequency Laboratories have demonstrated a reliable reception range of at least 150 miles for telephone weather broadcasts from Hadley Field, N. J. In the course of a number of flights, the Hadley beacon and weather services have been heard from Boonton, N. J., to Washington, D. C.

Similar ranges of reception of beacon and weather transmissions in different parts of the country have been observed in the course of a test flight from Boonton to Los Angeles and return.

Perhaps the most severe tests of the receiver by operators relatively unskilled in radio have been made on airplanes of the National Air Transport, Inc., in flights over the New York-Chicago air mail route. In the course of tests under all sorts of

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night and day flying conditions a reliable range of 150 miles on weather and beacon service has been reported. In a number of cases satisfactory beacon service has been obtained at a distance of 250 miles over mountainous territory. The airplanes used in this service are powered by Liberty motors with battery ignition. In such planes the level of ignition noise cannot readily be reduced to the negligible amount attainable in magneto ignited air-cooled motors.

Obviously any highly sensitive radio receiver imposes severe demands on the degree of ignition shielding. On planes equipped with battery ignition systems it is possible by careful shielding to reduce the disturbance for the receiver sensitivity herein described to a satisfactory low level. On planes using air-cooled motors with magneto ignition we have found that ignition disturbances may be completely eliminated by suitable shielding. Such shielding includes complete covering of magneto terminal blocks and all low-tension and high-tension wiring. Shielded spark plugs or their equivalent must be used.

In order to determine the relative contributions of the various parts of the ignition system to the general ignition disturbance a simple test has been devised. A receiver of the type described above is calibrated for sensitivity in microvolts against the position of the volume control rheostat. This receiver is then installed in a plane with a completely shielded ignition system. The test consists in removing various elements of the shielding and observing the points to which the receiver sensitivity must be reduced in order to make the interference just audible. Results of this test for a particular installation in an airplane equipped with a single Wright Whirlwind motor (Type J-5) are exhibited in Fig. 8, which shows for various amounts of shielding the corresponding maximum receiver sensitivities permissible with negligible ignition interference. These data may be translated directly into attainable reception range if the law of attenuation is known for wave propagation between ground and plane. Of course, information of the sort given in Fig. 8 must be applied with caution to other installations, where the various elements of the ignition system may be differently disposed with respect to the antenna; the only points of general interest are the relative magnitudes of the various factors contributing to the total disturbance. Of particular interest in Fig. 8 is the effect of shielded spark plugs in reducing ignition disturbance; they permit the use

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of a receiver forty times as sensitive as may be used with unshielded plugs.

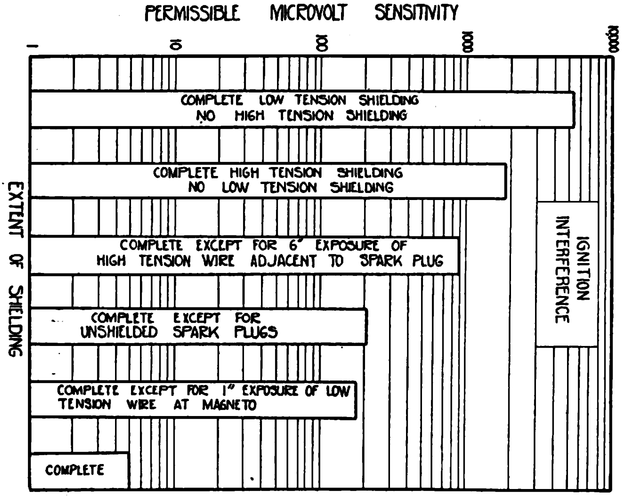

Fig. 8—Analysis of Ignition Disturbance from Wright Whirlwind Motor; Relation between Permissible Receiver Sensitivity and Degree of Shielding.

In conclusion the author wishes to acknowledge his indebtedness to Mr. W. D. Loughlin of the Engineering Division of Radio Frequency Laboratories for the structural design of the receiver.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

CIRCUIT ANALYSIS APPLIED TO THE SCREEN-GRID TUBE*

By

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Summary—*General radio-frequency circuit theory is discussed in this paper. The theoretical work and discussion is divided into two parts, amplification and stability.*

General amplification equations for impedance and transformer coupling, using an untuned primary whose period is above the highest frequency considered, are derived and discussed for the case of a screen-grid tube such as Cunningham type CX-322.

Feedback through the mutual capacity plate to grid capacity is also considered. A general expression for the limit of stable amplification per stage is inferred for n stages from the expressions found for one and two stages. This general expression $Av. < \sqrt{2gm/nwc_0}$ *is in terms of the mutual conductance, total grid to plate capacity, frequency and number of stages.*

CIRCUIT analysis is very simple with the perfect screen-grid tube. The mutual capacity between the input and output circuits is eliminated by the screening action of the fourth element, the screen grid. All other tube capacities are merely circuit constants. The plate impedance is infinite but the mutual conductance is finite. Under operating conditions the plate current is independent of plate voltage. The alternating component of the plate current is a function of only one tube constant, the mutual conductance. As the tube impedance is infinite the amplification is simply the product of the mutual conductance and external impedance.[1]

The mutual capacity, the control grid to plate capacity, of commerical screen-grid tubes such as Cunningham type CX-322 has been reduced to as small a value as is practical, less than 0.025 $\mu\mu$f. Although this capacity is very small it is large enough to affect the stability of an efficient radio-frequency amplifier at the higher broadcast frequencies particularly if care is not taken in the shielding. Neutralization of the mutual capacity is not necessary as stable amplification, four or five times that given

* Dewey decimal classification: R1320. Original manuscript received by the Institute, October 22, 1928.

[1] A. W. Hull and N. H. Williams, Characteristics of Shielded-Grid Pliotrons, *Phys. Rev.*, **27**, 432–438; April, 1926.

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by circuits using three-element tubes, is obtainable. The plate impedance is not infinite but has a value of about 800,000 ohms under operating conditions. The amplification, because of this finite plate impedance, is only approximately equal to the mutual conductance times the external impedance.

Hull[2] has discussed the circuit theory of the perfect screen-grid tube using impedance coupling. He made the assumption that the amplification is equal to the product of the mutual conductance and external impedance. An external plate impedance of around 150,000 ohms may be obtained in the broadcast band. This is appreciable compared with about 800,000 ohms plate resistance and its effect is to reduce the mutual conductance of the tube and load to some value less than the mutual conductance of the tube alone. The grid resistor across the tuned impedance also reduces its value. This effect is usually small unless the grid resistor has some value less than one megohm. Hull[2] found no indication of regeneration in making his overall amplification measurements as his calculated and measured values of amplification agreed. His measured amplification should have been about fifteen per cent lower than his calculated value as he assumed the amplification was equal to the product of the mutual conductance and external impedance. The regeneration factor for his circuit was also about fifteen per cent so that this would just about make up for the reduction of mutual conductance.

The purpose of this article is to consider the general theory of radio-frequency amplifiers and to apply this theory to the commercial screen-grid tubes. Feedback due to the mutual capacity, control grid to plate capacity, is considered in the general theory for n stages, and an expression for the limit of stable amplification is found. This paper will only cover, for steady conditions the circuits using impedance coupling, or transformer coupling with an untuned primary whose natural period is above the highest frequency considered.

Transformer Coupling

Where

μ is the amplification factor of the tube

r_p is the internal resistance of the tube

[2] A. W. Hull, Measurements of High Frequency Amplification with the Shielded-Grid Pliotron, *Phys. Rev.*, 27, 439–454; April, 1926.

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L_1 is the primary inductance
L_2 is the secondary inductance
M is the mutual inductance
C_2 is the secondary capacity
R_1 is the primary resistance
R_2 is the secondary resistance

Fig. 1 shows the equivalent circuit of a three-element tube and transformer. The screen-grid tube with a radio-frequency transformer may be represented by the same equivalent circuit

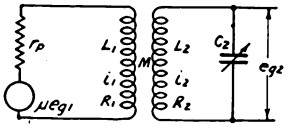

Fig. 1

if the d.c. bias of the screen grid is constant and if there is no external radio-frequency impedance in the screen-grid circuit. The equivalent values of μ and r_p vary with the d.c. voltages, but have definite values as soon as all the d.c. conditions are specified.

The voltage amplification Av. defined as the ratio of eg_2 to eg_1, is

$$\text{Av.} = \frac{\mu \quad \underline{|\delta}}{\omega C_2\left[\dfrac{Z_1{}^2 Z_2{}^2}{x^2{}_m} + x_m{}^2 - 2(x_2 x_1 - r_p R_2)\right]^{1/2}} \tag{1}$$

where

$$z_1 = \sqrt{r_p{}^2 + x_1{}^2}$$

$$z_2 = \sqrt{R_2{}^2 + x_2{}^2}$$

$$x_m = \omega m$$

$$\delta = \tan^{-1} \frac{r_p x_2 + R_2 x_1}{x_2 x_1 - x_m{}^2 - R_2 r_p}.$$

(For the derivation of the equations in this section see Appendix A.).

Equation (1) may be used to calculate the resonance curve. If we are interested only in the amplification at resonance simpler formulas may be derived. When any two of the three reactances X_1, X_2, or WM are varied so as to make I_2 a maxi-

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mum, the maximum possible amplification for a given coil and tube is obtained.

The value of this amplification is

$$\text{Av.} = \frac{1}{2} \frac{\omega L_2}{\sqrt{R_2}} \frac{\mu}{\sqrt{r_p}} \tag{2}$$

When the secondary reactance X_2 alone is varied to make I_2 a maximum the amplification is

$$\text{Av.} = \frac{\mu}{r_p} \frac{\omega L_2}{R_2\left[1+\dfrac{x_m^2}{R_2 r_p}\right]} x_m. \tag{3}$$

When X_2 is varied the frequency at which the circuit will resonate may be found from (7), Appendix A.

$$\left[r_p^2+\omega^2 L_1^2\right]\left[\omega L_2 - \frac{1}{\omega C_2}\right] = \omega L_1 \omega^2 M^2. \tag{4}$$

If the primary is open rp is almost infinite and as $\omega L_1 \omega^2 M^2$ is finite, the value of $(\omega L_2 - 1/\omega C_2)$ must approach zero so the resonant frequency is determined by the values of the secondary capacity and inductance.

$$\omega_r = \frac{1}{\sqrt{L_2 C_2}} \tag{5}$$

where ω_r is 2π times the resonant frequency.

When r_p is not infinite ω_r becomes

$$\omega_r^2 = -\omega_0^2\left[\frac{r_p^2 L_2 C_2}{2L_1^2} - \frac{1}{2}\right] + \sqrt{\left(\frac{r_p^2 L_2 C_2}{2L_1^2} - \frac{1}{2}\right)^2 \omega_0^4 + \frac{\omega_0^2 r_p^2}{L_1^2}} \tag{6}$$

where

$$\omega_0^2 = \frac{1}{C_2 L_2 (1-K^2)}$$

K = coefficient of magnetic coupling.

If r_p in (6) is made zero by shorting the primary on itself ω_r becomes

$$\omega^2_r = \tfrac{1}{2}\omega_0^2 + \tfrac{1}{2}\omega_0^2 = \omega_0^2 \tag{7}$$

The values of ω_r given by (5), (6), and (7) are independent of the secondary resistance. Under the usual circuit conditions

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if the value of r_p is small, the coupling is weak so that (5) may be used without introducing much error.

Impedance Coupling

Impedance coupling will be considered next. This should be the equivalent of transformer coupling when both inductances are equal and the coefficient of magnetic coupling is unity.

Fig. 2-A shows the circuit using impedance coupling. When

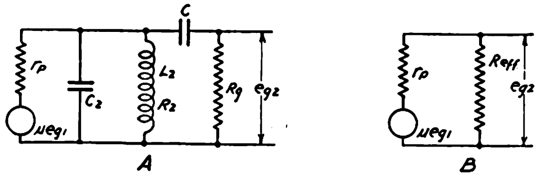

Fig. 2

resonance is defined as the frequency at which the circuit has zero reactance the combination L_2, R_2, C_2 acts a resistance whose value is

$$R \text{ equivalent} = \frac{L_2}{R_2 C_2} = \frac{\omega^2 L_2^2}{R_2} \tag{8}$$

If the coupling condenser C is large enough to have a small reactance at the frequency considered, Fig. 2-A may be replaced by Fig. 2-B. The effect of Rg is to add a resistance to the secondary resistance R_2 and the value of this added resistance is given by

$$R = \frac{\omega^2 L_2^2}{R_g} \cdot \tag{9}$$

Calling the total effective resistance of the tuned circuit $R^1{}_2$ the value of R eff. is now

$$R_{eff} = \frac{\omega^2 L_2^2}{R_2 + \dfrac{\omega^2 L_2^2}{R_g}} = \frac{\omega^2 L_2^2}{R_2'} \tag{10}$$

$$i_p = \frac{\mu e_{g1}}{r_p + R_{eff}} \tag{11}$$

$$e_{g2} = \frac{\mu e_{g1} R_{eff}}{r_p + R_{eff}} \tag{12}$$

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$$\text{Av.} = \frac{\mu}{p}\,\omega L_2 \frac{\omega L_2}{R_2'\left(1+\dfrac{\omega^2 L_2{}^2}{R_2' r_p}\right)} \tag{13}$$

Equation (13) is of the same form as (3) with ωL_2 in (13) replacing ωM of (3). $R^1{}_2$ of (13) includes the effect of the grid resistor R_g being shunted across the tuned circuit. The above expression for Av. will only give the amplification at resonance. The response at any other frequency may be found from (14) derived in Appendix B.

$$\text{Av.} = \frac{\mu}{r_p}\,\frac{\omega^2 L_2{}^2 + R_2'^2}{\sqrt{[C_2\omega(\omega^2 L_2{}^2 + R_2'^2) - \omega L_2]^2 + \left(\dfrac{\omega^2 L_2{}^2}{r_p} + R_2'\right)^2}} \tag{14}$$

Hull gives the same value of impedance with the exception that his term does not contain $\omega^2 L_2{}^2/r_p$. With a perfect screen-grid tube this term would be zero because of the high plate impedance, but with the commercial screen-grid tubes it should be considered as its value is appreciable.

The amplification neglecting $R_2{}^{12}$ where it is added to $\omega^2 L_2{}^2$ is

$$\text{Av.} = \frac{\mu}{r_p}\,\frac{\omega^2 L_2{}^2}{\sqrt{[\omega L_2(\omega^2 L_2 C_2 - 1)]^2 + \left[\dfrac{\omega^2 L_2{}^2}{r_p} + R_2'\right]^2}} \tag{15}$$

at resonance $\omega^2 L_2 C_2 = 1$ and (15) becomes

$$\text{Av.} = \frac{\mu}{r_p}\,\frac{\omega^2 L_2{}^2}{R_2' + \dfrac{\omega^2 L_2{}^2}{r_p}} \tag{16}$$

which is equivalent to (13).

Discussion of the Amplification Equations

In the usual multistage set it is not convenient to vary L_1. A maximum value of secondary current may be obtained by varying any two of the three reactances x_1, x_2, and x_m. Varying both x_m and x_2, the amplification obtained is given by Eq. (2). This will give the maximum amplification obtainable. The selectivity given by

$$S = \frac{\omega^2 L_2{}^2}{R_2'} \tag{17}$$

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(where R_2' includes the secondary resistance R_2 and the resistance transferred from the primary ω^2M^2/r_p) is only one-half that of the coil alone.

The condition for optimum coupling is given by (9), Appendix A, as

$$x_m{}^2 = R_2r_p + x_1x_2$$

As x_2 is small this is approximately

$$x_m{}^2 = R_2r_p \qquad (18)$$

With the screen-grid tube the plate resistance r_p is around 800,000 ohms under operating conditions. R_2 will be at least 5 ohms so (18) cannot be satisfied. Eq. (2) will not apply to this tube. To obtain the maximum amplification impedance coupling should be used.

As the condition for optimum coupling cannot be realized the selectivity of the tuned circuit will always be greater than one-half that of the coil alone.

The value of x_2 is varied so (3) will apply to this tube using transformer coupling. Almost as much amplification may be obtained with a properly designed transformer coupled circuit as with an impedance coupled circuit. The primary inductance may be slightly larger than the secondary inductance. If the primary is made small physically by winding it with small wire it may be coupled rather closely magnetically to the secondary without adding excessive dielectric losses to the tuned circuit.

There is some gain in selectivity by using a carefully designed transformer over that obtained by using impedance coupling. The effective resistance added to the secondary when a transformer is used is ω^2M^2/r_p. The resistance added by impedance coupling is $\omega^2L_2{}^2/r_p$ plus a term $\omega^2L_2{}^2/R_g$. The resistance added to the secondary by transformer coupling is less than that added by impedance coupling, and if the transformer is well designed the transformer coupled circuit will have better selectivity than the impedance coupled circuit.

All the amplification equations show that the amplification will decrease with a decrease of mutual conductance. Fig. 3 shows μ, r_p and gm plotted against the screen-grid voltage, all other voltages being left constant. The screen-grid voltage may be used as a volume control. The value of r_p will increase as gm is decreased so the selectivity will increase as the volume is

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decreased. Tuning of the circuits will not be affected by the volume control.

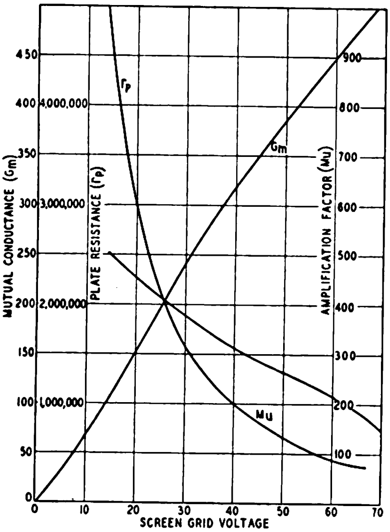

Fig. 3—CX-322 Screen-Grid Connection. Plate resistance, mutual conductance, and mu vs. screen-grid voltage. $Ec_1 = 1.5$ volts, $Eb = 135$ volts.

Stability

The control grid to plate capacity couples the input and output circuits in an amplifier. The output circuit contains a source of energy and will react on the input circuit because of the mutual reactance. The voltage fed back from the output to the input may be either in phase or out of phase with the input voltage. If the amount of energy fed back is enough to supply the input losses the circuit will oscillate. The circuit conditions necessary to cause oscillation through the mutual tube capacity depend upon the number of stages. The amplification increases geometrically and the capacity coupling the input and output decreases arithmetically so that a stage is reached where the output will supply enough energy to supply the input losses causing the amplifier to oscillate.

Hull[3] has discussed stability by assuming that maximum regeneration will be obtained when all the circuits are in exact

[3] loc. cit.

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resonance. He does not consider n stages, but derives results for only two stages.

Any discussion of stability should make no assumptions as to values of the phase angles. The theory should also show what factor multiplies the amplification value if oscillations do not occur. The final results in order to be general should be for n stages.

Beatty[4] has found the circuit conditions that will cause one stage to oscillate. His limit as to the value of certain circuit constants that will cause oscillations to occur is found graphically without any mathematical proof. This article will show that his limit is correct, and will in addition carry out the work to two stages. A general law for n stages may be inferred from these results.

Fig. 4 shows the circuit that will be considered. Any transformer coupled circuit may be reduced to this circuit so the results will be general. The coils are assumed to be similar. The

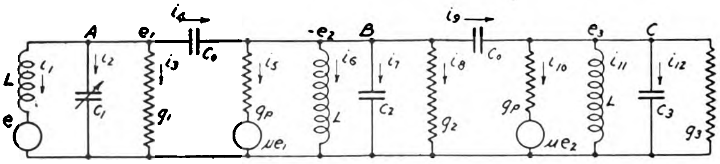

Fig. 4

resistances in the coils have been replaced with conductances g_1, g_2, g_3 in parallel with the tuned circuits. The value of this conductance is

$$gc = \frac{R_2}{\omega^2 L^2}. \tag{19}$$

The plate to filament conductance is g_p. A voltage e is induced in the circuit. The ratio of the voltage e_1 and e is

$$\frac{e_1}{e} = \frac{\omega L}{R} \quad \text{or} \quad \frac{1}{gc\omega L}. \tag{20}$$

For the derivation of equations in this section see Appendix C.

[4] R. T. Beatty, The Stability of the Tuned-Grid Tuned-Plate, H F Amplifier, *Exper. Wireless and Wireless Engineer*, V, 3; January, 1928.

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Beatty[5] by means of (1C) and (2C), Appendix C, without the last term $(-e_3 j\omega C_0)$ in Eq. (2C), found the following relation between e_2 and e.

$$\frac{e_2}{e} = jAF \tag{21}$$

where

$$\frac{1}{F} = (1 + j \tan \theta_1)(1 + j \tan \theta_2) + jH \tag{22}$$

$$H = \frac{gmC_0\omega}{gc(gp + gc)} \tag{23}$$

$$A = \frac{gm}{\omega L_1 gc(gp + gc)} \cdot \tag{24}$$

The factor F will be infinity when

$$-jH = (1 + j \tan \theta_1)(1 + j \tan \theta_2) \tag{25}$$

H is a circuit constant depending on ω, gm, C_0 and the coils.

Beatty[6] shows graphically that $1/F$ is the distance between H and a parabola. When H is 2 this vector will be zero and the circuit will oscillate.

The smallest value of H that will cause the circuit to oscillate can be shown mathematically to be 2 as follows:

$$-jH = (1 + jx)(1 + jy) \tag{26}$$

or

$$-jH = 1 + jy + jx - xy. \tag{27}$$

Equating reals and imaginaries,

$$xy = 1$$

$$-H = y + x. \tag{29}$$

From (28) x and y must have the same sign.

From (29) x and y must be negative.

This means that the phase angles are negative or both circuits will have inductive reactance.

From (28) $x = 1/y$.

Substituting this in (29)

[5] loc. cit.
[6] loc. cit.

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$$y+\frac{1}{y}+H=0 \quad \text{or} \quad y^2+1+Hy=0 \tag{30}$$

$$y=\frac{-H\pm\sqrt{H^2-4}}{2} \tag{31}$$

y must be real so that H^2 must be equal to or greater than 4, hence the smallest value that H may have that will cause the circuit to oscillate is 2. When H is 2 each circuit must have its effective resistance equal to the effective inductance. If H is larger than 2 this restriction no longer holds. As we are only interested in the smallest value of H that will cause oscillations to occur the values of the phase angles when H is greater than 2 are of no interest in this discussion.

If H is smaller than 2 the regeneration factor for one stage may be found as explained in Beatty's[7] article. This phase of the question will not be further discussed here.

The relation between e_3 and e_1 is from (6) Appendix C.

$$\frac{e_3}{e}=\frac{gm^2}{j\omega L_1 g_1 g_4 g_5}\,\frac{1}{A+B} \tag{32}$$

where

$$A=(1+j\tan\theta_1)(1+j\tan\theta_2)(1+j\tan\theta_3) \tag{33}$$

$$B=j\omega C_0 gm\left(\frac{1}{g_4\,g_5}+\frac{1}{g_4\,g_1}\right)$$

$$-\omega C_0 gm\left(\frac{\tan\theta_1}{g_4\,g_5}+\frac{\tan\theta_3}{g_4\,g_1}\right). \tag{34}$$

Let $g=g_1=g_4=g_5$ as would be the case in a multistage set. Eq. (34) becomes

$$B=j\frac{2\omega C_0 gm}{g^2}-\frac{\omega C_0 gm}{g^2}(\tan\theta_1+\tan\theta_3) \tag{35}$$

The quantity $\omega C_0 gm^2/g^2$ is a circuit constant. Denote its value by H'. Writing x for $\tan\theta_1$, y for $\tan\theta_2$, z for $\tan\theta_3$ we have the following expression:

$$A+B=(1+jx)(1+jy)(1+jz)+2jH'-H'(x+y).$$

The condition for the circuit to oscillate is that $A+B=0$.

[7] loc. cit.

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In the single stage the conditions to cause oscillations were that H should be 2 and the phase angles negative and 45 deg. Assuming that the phase angles will be equal and the tangents equal to -1, we find that H' for two stages is equal to 1.

TABLE I

X	Y	Z	H'
	0	(Impossible as H' is imaginary)	
−4	−0.25	−4	2.125
−2	−0.50	−2	1.25
−1	−0.75	−3	1.25
−1.33	−0.75	−1.33	1.04
−1	−0.78	−2	1.11
−1	−0.84	−1.5	1.04
−1	−1	−1	1.00
−0.5	−1.44	−1	1.11
−0.667	−1.5	−0.667	1.08
−0.5	−2.0	0.5	1.25
−0.5	−2.6	0.3	1.50
−0.333	−3.0	0.333	1.67
−0.0	−3.0	−1	2.0
−0.1	−3.0	0	2.0
−0.25	−4.0	−0.25	2.125
	+1	(Impossible as H' must be −1.)	

Table I shows how the value of H' necessary to cause the circuit to oscillate varies with the phase angles. From this table it is seen that any value of H equal to or greater than 1 will cause the circuit to oscillate. Oscillations cannot occur if H is less than 1. From the table y must always be some negative value. When either x or z is made zero, H is 2, the same value as found for one stage. The phase angles are different, which is to be expected. Assume one stage was oscillating and another stage was added. If the phase angle was zero this would merely act as an amplifier, and to cause the whole circuit to oscillate it would be necessary to change the tuning on the first two condensers.

The results to be general should be expressed in terms of Av. Divide both the numerator and denominator of (13) by $\omega^2 L^2$

$$\text{Av.} = \frac{\mu}{r_p} \frac{1}{\dfrac{R_2}{\omega^2 L^2} + \dfrac{1}{r_p}} = \frac{gm}{gc + gp} = \frac{gm}{g}. \tag{36}$$

Refer to (32). As was previously mentioned, $1/j\omega L g_1$ is the ratio between e_1 and e. The remaining part of (32) $gm^2/g_4 \cdot g_5(A+B)$ is the ratio of the voltage e_3 to e_1. If regeneration is not taken into account, $gm^2/g_4 \cdot g_5$ is the amplification of two stages as the amplification of one stage is gm/g as was shown in (66). When the value of the vector $1/(A+B)$ is known the total amplification considering reaction is the amplification per stage squared times the regeneration factor $1/(A+B)$.

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For one stage we found

$$H=2=\frac{\omega C_0 gm}{g^2} \tag{37}$$

for two stages

$$H'=1=\frac{\omega C_0 gm}{g^2} \tag{38}$$

multiply (38) by 2

$$2=2\omega C_0 gm/g^2 \tag{39}$$

or

$$2=\frac{n\omega C_0 gm}{g^2}=\frac{n\omega C_0 \text{Av.}^2}{gm} \tag{40}$$

where n is the number of stages or the number of tuned circuits minus 1.

$$\text{Av.}^2=\frac{2gm}{n\omega C_0}\cdot \tag{41}$$

For the circuit to be stable,

$$\text{Av.}<\sqrt{\frac{2gm}{n\omega C_0}}\cdot \tag{42}$$

Hull[2] ound an expression for two stages,

$$\text{Av.}<\sqrt{\frac{gm}{\omega C_0}+1}\ +1.$$

When n is 2, Eq. (42) becomes

$$\text{Av.}<\sqrt{\frac{gm}{\omega C_0}}\cdot \tag{43}$$

Hull's expression agrees with this expression as to the circuit values and differs in adding 1 under the square root sign and 1 to the whole expression. As he assumed exact resonance there is no way of telling from his equations whether there would be any feedback if the circuit did not oscillate. In other words according to his results there would be no regeneration until the point of oscillation was reached, when the circuit would suddenly start to oscillate.

Eq. (42) shows the importance of shielding the input from the output in using a tube with a low control grid to plate capacity

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if stability is to be obtained without neutralization or adding losses.

The tube should be operated so that the value of gm is as high as is consistent with good tube life as the stable value of Av. in (42) will be raised with an increase of gm.

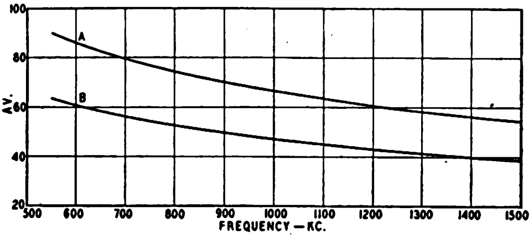

Fig. 5—Limit of Stable Amplification per Stage for Type CX-322 Tubes. $gm=350$ micro-mhos, $Co=0.25$ $\mu\mu$f, *A*—one-stage amplifier, *B*—two-stage amplifier.

Fig. 5 shows the limit of stable amplification plotted for a CX-322 tube. The value of C_0 is taken as 0.025 $\mu\mu$f. This value is the extreme limit and most of the tubes have less than 0.02 $\mu\mu$f control grid to plate capacity. The value of gm is taken as 350×10^{-6} mhos, which is an average value.

Eq. (42) is general and will apply to any tube where the only feedback is through the mutual capacity. Most three-element tube circuits are stabilized in some manner. Limits of stable amplification using three-element tubes are of more theoretical than of practical interest and will not be given here.

Note: Since this article was written Rockwood and Thompson[8] published a formula for the maximum load impedance which may be used with a tube without oscillations occurring.

$$R_p=\frac{1.41r_p}{r_p(2\pi fC_{p-g}gm)^{1/2}-1.41}.$$

This expression is the same as (42) when n is 1, as can be seen from the following algebraic transformation.

Clearing of fractions,

$$Rprp\sqrt{\omega C_0 gm}-\sqrt{2}Rp=\sqrt{2}r_p$$

$$\sqrt{2}(Rp+rp)=Rprp\sqrt{\omega C_0 gm}$$

[8] Rockwood and Thompson, Discussion on Application of the Four Electrode Receiving Tube (UX-222), *Proc.* of the Radio Club of America, 5, 69; Jun e, 1928.

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$$\sqrt{29M}(Rp+rp)=Rprpgm\sqrt{\omega C_0}$$

$$\frac{\mu}{Rp+rp}R_p=\sqrt{\frac{2gm}{\omega C_0}}$$

$$\text{Av.}=\frac{\mu}{Rp+rp}R_p=\sqrt{\frac{2gm}{\omega C_0}}$$

which is the same as Eq. (42) when n is 1.

Appendix A

Derivations of Equations for Fig. 1

$$I_2=\frac{j\omega M\mu eg_1}{r_pR_2+\omega^2M^2-x_1x_2+j(r_px_2+\omega L_1R_2)} \tag{1}$$

R_1 has been neglected because it is small compared to r_p where

$$x_2=\omega L_2-\frac{1}{\omega C_2}$$

$$x_1=\omega L_1$$

$$eg_2=\frac{I_2}{j\omega C_2} \tag{2}$$

$$\text{Av.}=\frac{eg_2}{eg_1}=\frac{\mu\omega M}{\omega C_2\{r_pR_2+\omega^2M^2-x_1x_2)+j(r_px_2+x_1R_2)} \tag{3}$$

The absolute value of Av. is

$$\text{Av.}=\frac{\mu \;\underline{|\delta}}{\omega C_2\left[\dfrac{Z_1{}^2Z_2{}^2}{\omega^2M^2}+\omega^2M^2-2(x_1x_2-r_pR_2)\right]^{1/2}} \tag{4}$$

$$Z_{12}{}^2=\frac{eg_1{}^2}{I_2{}^2}=\frac{Z_1{}^2Z_2{}^2+\omega^4m^4-2\omega^2M^2(x_1x_2-r_pR_2)}{\omega^2M^2} \tag{5}$$

The value of I_2 will be a maximum when Z_{12} is a minimum. To find when I_2 will be a maximum, as any of the reactances are varied, it is only necessary to differentiate $Z_{12}{}^2$ with respect to that reactance and to set the derivate thus found equal to zero and to solve for the desired reactance in terms of the other impedances.

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Doing this we find that I_2 will be maximum when

$$\omega L_1=\frac{\omega^2M^2x_2}{R_2^2+x_2^2} \tag{6}$$

$$x_2=\frac{\omega^2M^2x_1}{r_p^2+x_1^2} \tag{7}$$

$$\omega^2M^2=(r_p^2+x_1^2)[R_2^2+x_2^2). \tag{8}$$

Relations to satisfy (8), (9), and (10) may be found by equating values of ω^2M^2 from these equations. Doing this we find

$$\frac{x_2}{R_2}=\frac{x_1}{r_p}. \tag{9}$$

When (9) is satisfied ω^2M^2 from (8) becomes

$$\omega^2M^2=R_2r_p+x_1x_2=\frac{R_2}{r_p}(r_p^2+x_1^2) \tag{10}$$

If we satisfy any two of (6), (7), and (8), the third will also be satisfied. It is not usually convenient to vary x_1. Varying both x_2 and ωM, Z_{12} becomes

$$Z_{12}=2\sqrt{R_2r_p}\frac{\sqrt{r_p^2+\omega^2L_1^2}}{\omega L_1+jr_p} \tag{11}$$

I_2 from Eq. (1) becomes

$$I_2=\frac{1}{2}\frac{\mu}{\sqrt{R_2r_p}}\frac{x_1+jr_p}{\sqrt{r_p^2+x_1^2}}. \tag{12}$$

If r_p is large compared to ωL_1,

$$\text{Av.}=\frac{1}{2}\frac{\omega L_2}{\sqrt{R_2}}\frac{\mu}{\sqrt{r_p}}. \tag{13}$$

If only x_2 is varied Z_{12} is

$$Z_{12}=\frac{R_2\sqrt{\omega^2L_1^2+r_p^2+\dfrac{\omega^2M^2r_p}{\omega L_1+jr_p}}}{\omega^2M^2}. \tag{14}$$

If r_p is large compared to (ωL_1) Av. is

$$\text{Av.}=\frac{\mu}{r_p}\frac{\omega L_2}{R_2}\frac{\omega M}{1+\dfrac{\omega^2M^2}{R_2r_p}}. \tag{15}$$

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Appendix B

General Equations for Fig. 2-A.

Let Z be the impedance of the tuned circuit.

$$Z=\frac{\omega L-jR_2}{R_2'\omega C_2+j(\omega^2L_2C_2-1)} \tag{1}$$

where R_2' includes the effect of the grid resistor.

Let Z_1 be the total impedance of the circuit.

$$Z_1=Z+r_p=\frac{r_pR_2'\omega C_2+j(\omega^2L_2C_2-1)+\omega L-jR_2'}{R_2'\omega C_2+j(\omega^2L_2C_2-1)} \tag{2}$$

$$i_p=\frac{\mu eg_1}{Z_1} \tag{3}$$

$$e_p \text{ or } eg_2=\frac{\mu eg_1}{Z_1}Z \tag{4}$$

$$\text{Av.}=\frac{eg_2}{eg_1}=\mu\frac{\dfrac{\omega L_2-jR_2'}{R_2'\omega C_2+j(\omega^2L_2C_2-1)}}{\dfrac{R_2'r_p\omega C_2+\omega L_2+j(r_p\omega^2L_2C_2-r_p-R_2')}{R_2'\omega C_2+j(\omega^2L_2C_2-1)}} \tag{5}$$

The absolute value of Av., neglecting the phase angle and R_2'/r_p, is

$$\text{Av.}=\frac{\mu}{r_p}\frac{\omega^2L_2^2+R_2'^2}{\sqrt{[C_2\omega(\omega^2L_2^2+R_2'^2)-\omega L_2]^2+\left[\dfrac{\omega^2L_2^2}{r_p}+R_2'\right]^2}} \tag{6}$$

Appendix C

Analysis of Fig. 4

Applying Kirschoff's first law to the points A, B, and C, we have

$$i_1+i_2+i_3+i_4=0 \tag{1}$$

$$\frac{e_1-e}{j\omega L}+e_1j\omega C_1+e_1g_1+[e_1-(e_2)]j\omega C_0=0 \tag{1a}$$

$$e_1\left[g_1+j\left\{\omega(C_1+C_0)-\frac{1}{\omega L}\right]-\frac{e}{j\omega L}+e_2j\omega C_0=0 \tag{1b}$$

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Let $\tan\theta_1$ be $\dfrac{\omega(C_1+C_0)-\dfrac{1}{\omega L}}{g_1}$; then Eq. (1B) becomes

$$e_1g_1(1+j\tan\theta_1)-\frac{e}{j\omega L}+e_2j\omega C_0=0 \qquad (1c)$$

$$-i_4+i_5+i_6+i_7+i_8+i_9=0 \qquad (2)$$

$$-(e_1+e_2)j\omega C_0+(-e_2+\mu e_1)gp-\frac{e_2}{j\omega L}-e_2gc-(e_2+e_3)j\omega C_0=0 \qquad (2a)$$

$$-e_2\left(j\omega(C_2+2C_0)-\frac{j}{\omega L}+gp+g_2\right)-e_1j\omega C_0+gme_1-e_3j\omega C_0=0 \qquad (2b)$$

$$gm=\mu gp$$

$$\text{Let}\quad gp+g_2=g_4$$

ωC_0 may be neglected in comparison with gm. Eq. (2B) then becomes

$$-e_2g_4(1+j\tan\theta_2)+e_1gm-e_3j\omega C_0=0 \qquad (2c)$$

$$-i_9+i_{10}+i_{11}+i_{12}+i_{13}=0 \qquad (3)$$

$$(e_2+e_3)j\omega C_0+(e_3-\mu e_2)g_p+\frac{e_3}{j\omega L}+e_3j\omega C_3+e_3g_3=0 \qquad (3a)$$

$$e_3\left[j\omega(C_0+C_3)-\frac{j}{\omega L}+gp+g_3\right]+e_2j\omega C_0-e_2gm=0 \qquad (3b)$$

$$e_3g_5(1+j+\tan\theta_3)-e_2gm=0 \qquad (3c)$$

Multiply (1C) by gm and (2C) by $g_1(1+j\tan\theta_1)$.

Subtract (2C) from (1C) after multiplying by the above quantities and we have

$$e_2\{j\omega C_0gm+g_4g_1(1+j\tan\theta_1)(1+j\tan\theta_2)\}$$
$$+e_3j\omega C_0g_1gm(1+j\tan\theta_1)=\frac{egm}{j\omega L} \qquad (4)$$

Eliminating e_2 between (4) and (3C) we have

$$e_3(j\omega C_0gmg_1(1+j\tan\theta_1)+j\omega C_0gmg_5(1+j\tan\theta_3)$$

$$+g_5g_4g_1(1+j\tan\theta_1)(1+j\tan\theta_2)(1+j\tan\theta_3)=\frac{eg^2m}{j\omega L} \qquad (5)$$

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$$\frac{e_3}{e} = \frac{gm^2}{j\omega L_1 g_1 \; g_4 \; g_5} \cdot \frac{1}{A+B} \tag{6}$$

where

$$A = (1+j \tan \theta_1)(1+j \tan \theta_2)(1+j \tan \theta_3) \tag{7}$$

$$B = j\omega C_0 gm\left(\frac{1}{g_4 \; g_5} + \frac{1}{g_4 \; g_1}\right) - \omega C_0 gm\left(\frac{\tan \theta_2}{g_4 \; g_5} + \frac{\tan \theta_3}{g_4 \; g_1}\right) \tag{8}$$

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

THE EFFECT OF REGENERATION ON THE RECEIVED SIGNAL STRENGTH*

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Summary—*It is the purpose of this paper to give a theory of the effect of regeneration using the solution of a non-linear differential equation, and to present experimental verification of the theory.*

It is shown that: (a) as a first approximation, detection has no effect on the radio-frequency grid voltage developed under the influence of an incoming signal, (b) the amplification obtained through regeneration equals the two-thirds power of the ratio of the "grid space" to the amplitude obtained with zero regeneration. It is apparent from (b) that much greater gain is obtained through regeneration with weak signals than with strong signals.

The verification of the theory is made with a circuit arrangement operating at 500 cycles per second. Application is made to radio frequencies using this as a model of a high-frequency system, following a theorem for model systems which is stated.

THE considerable increase in signal strength obtainable through the use of regeneration is well known. It is also common knowledge that this increase in signal strength is considerably greater when the incoming signal is weak than when it is strong. It is the purpose of this paper to provide a non-linear theory of this effect of regeneration and the experimental verification thereof.

Suppose a triode system of one degree of freedom, as in Fig. 1 where L, C, r form the tuned circuit and the mutual induction M provides the regeneration.

Let further an emf, $E \sin \omega_1 t$, representing the "incoming signal," to be applied to the oscillatory system. Neglecting the grid current and calling the current in the LCr circuit i, and the deviation of the anode current from its steady value i_a and the alternating grid P.D. V_g, we have

$$L\frac{di}{dt}+ri+\frac{1}{C}\int idt-M\frac{di_a}{dt}=E\sin\omega_1 t \qquad (1)$$

$$\frac{1}{C}\int idt=v_g \qquad (2)$$

* Dewey decimal classification: R134.4. Corrected copy; reprinted from August, 1928, issue of the PROCEEDINGS. Paper read at the meeting of the International Union of Scientific Radio-telegraphy, October, 1927.

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In the "tuned grid" circuit, the variations of anode potential are usually small compared with the variations of the grid potential, so that as a first approximation, the anode current variation i_a is a function of the grid potential variation only, i.e.,

$$i_a = f\ (v_g) \tag{3}$$

In order to avoid complexity, we neglect the grid current and we therefore imagine a negative grid bias to be provided. Our system of coördinates in the i_a, v_g plane is therefore as shown in Fig. 2. This differs from the usual notation in so far as the zero point of the coördinate system is shifted towards the steady d.c. position round which the oscillations occur.

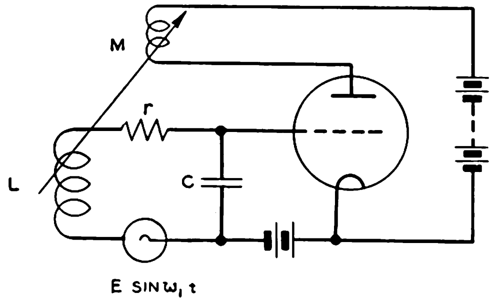

Fig. 1

We approximate equation (3), representing the curved plate current-grid voltage characteristic of the triode, by the cubic:

$$i_a = S_1 v_g + S_2 v_g^2 - S_3 v_g^3 \tag{3a}$$

where S_1 is the usual "mutual conductance" for infinitesimal grid-potential variations. S_2 and S_3 are further determined by the form of the characteristic. As the latter bends round both at the top and at the bottom we write in (3a) $-S_3 v_g^3$ instead of $+S_3 v_g^3$.

The elimination of i and i_a from (1), (2), and (3a) results in:

$$\frac{d^2 v_g}{dt^2} + \left\{ \left(\frac{r}{L} - \frac{MS_1}{LC} \right) - \frac{2MS_2}{LC} v_g + \frac{3MS_3}{LC} v_g^2 \right\} \frac{dv_g}{dt} + \omega_0^2 v_g = \omega_0^2 E \sin \omega_1 t. \tag{4}$$

where

$$\omega_0^2 = \frac{1}{LC} \cdot$$

Calling further

$$\frac{r}{L} - \frac{MS_1}{LC} = \alpha,$$

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$$\frac{MS_2}{LC}=\beta,$$

$$\frac{MS_3}{LC}=\gamma,$$

(4) becomes

$$\ddot{v}_g+(-\alpha-2\beta v_g+3\gamma v_g^2)\,\dot{v}_g+\omega_0^2 v_g=\omega_0^2 E\sin\omega_1 t. \qquad (4a)$$

which is a non-linear inhomogeneous differential equation of the second order with non-linear resistance terms. The usual elementary approximation of a linear characteristic would make $\beta=\gamma=0$ and for resonance (i.e. $\omega_0^2=\omega_1^2$) and critical regeneration (i.e. $\alpha=0$) the developing grid voltage v_g would become infinite. In order to obtain a satisfactory theory of the response of a regenerative triode system to an expressed emf a non-linear problem must therefore be solved.

The equation (4a) was fully considered in a former paper[1]; from the results obtained there it follows that the steady state solution in the neighborhood of resonance ($|\omega_0-\omega_1|\ll\omega_1$) can be written:

$$v_g=b\sin(\omega_1 t+\phi)$$

where the amplitude b of the resulting grid potential variation is given by:

$$b^2\left\{4(\omega_0-\omega_1)^2+(\alpha-\tfrac{3}{4}\gamma b^2)^2\right\}=\omega_0^2E^2 \qquad (5)$$

a cubic equation in b^2. It is further seen from (5) that the symmetrical term in (4a) (with β) (which determines the detection) has, in the first approximation here considered, no influence on the resulting v_g. *Therefore as a first approximation detection has no effect on the h.f. grid potential difference developing under the influence of an incoming "signal."*

Further if (5) is compared with the usual linear case as represented by

$$b^2\left\{4(\omega_0-\omega_1)^2+\alpha_0^2\right\}=\omega_0^2E^2,$$

it follows that in the non-linear case $(\alpha-\frac{3}{4}\gamma b^2)$ is substituted for α_0, i.e. the system behaves as if for the resistance r a new resistance r' were substituted of a value given by

$$r'=r-\frac{M}{C}\left(S_1-\frac{3}{4}b^2S_3\right) \qquad (6)$$

[1] *Phil. Mag.* **3**, 65, 1927.

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which therefore depends upon the amplitude b already present in the system.[2]

When the regeneration is pushed so far that

$$MS_1 > rC \tag{7}$$

it is seen from (6) that the first order differential resistance of the system becomes negative and therefore the system has the tendency to oscillate spontaneously. However, as was shown in the *Phil. Mag.* paper quoted above, the forced oscillations may suppress the development of the free oscillations. The phenomenon manifests itself through the presence of a "silent region" extending at both sides of resonance. As was shown by Professor Appleton[3] the width of this silent region is determined by the amplitude of the incoming signal. For strong signals this width is given by

$$\frac{\omega_0 - \omega_1}{\omega_0} = \pm E\sqrt{\frac{3\gamma}{\alpha}} \tag{8}$$

It is of interest to investigate the resultant grid amplitude b if (**a**), the system is tuned exactly to resonance, i.e.,

$$\omega_0 = \omega_1$$

and (**b**), it is brought on the verge of free oscillation, i.e., when

$$S_1 M = Cr. \tag{9}$$

For this case we at once obtain from (5)

$$\tfrac{3}{4}\gamma b^3 = \pm \omega_0 E, \tag{10}$$

which expression will now be considered in detail.

First, it follows from (10) that *the resulting grid amplitude is proportional to the cube root of the emf applied to the system.*

Further, it is easy to assign an approximate value to γ directly from the triode characteristic. Referring to (3a), (4a) and Fig. 2, and taking the symmetrical case for which $S_2 = 0$, it follows that a good approximation to S_3 can be found from the maximum and minimum value of the cubic (3a), for $i_a = i_{a\ max}$ when

$$v_g{}^2 = \frac{S_1}{3S_3}\,.$$

[2] This property of the non-linear system was first derived in 1920. See Balth. van der Pol, *Radio Review*, Nov., Dec., 1920.

[3] E. V. Appleton, *Proc.* Camb. Phil. Soc., **23**, 231, 1923.

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Calling, therefore, the grid voltage change necessary to bring the anode current from zero to its saturation value the "grid space" and designing it by V_{g0} (see Fig. 2), we obtain

$$V^2_{g0} = \frac{4S_1}{3S_3} \tag{11}$$

hence

$$\gamma = \frac{M}{LC} \cdot \frac{4S_1}{3V^2_{g0}} .$$

But, when the system has critical regeneration (9) obtains, i.e.,

$$M = \frac{Cr}{S_1}$$

hence

$$\gamma = \frac{r}{L} \cdot \frac{4}{3V^2_{g0}} .$$

When we further call v_{g1} the grid-voltage amplitude which would be obtained with no regeneration at all, (M = 0), or (which is the same thing) with reduced filament current, we find from (5)

$$v^2_{g1} = \frac{\omega^2_0 L^2}{r^2} \cdot E^2 ,$$

hence we obtain from (10):

$$b^3 = v_{g1} \cdot V^2_{g0}$$

or

$$b = \sqrt[3]{v_{g1} \cdot V^2_{g0}} \tag{12}$$

reading in words:

The grid amplitude developing in resonance and with critical regeneration equals the cube root of the product of the grid amplitude which would be obtained with no regeneration at all, and the square of the "grid space," as defined by Fig 2.

(12) can further be written:

$$\frac{b}{v_{g1}} = \left(\frac{V_{g0}}{v_{g1}}\right)^{2/3} \tag{13}$$

which means that: *the amplification obtained through regeneration equals the two-third power of the ratio of the "grid space" V_{g0} into the amplitude obtained with zero regeneration.* The much greater

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gain obtained through regeneration with small signals than with stronger ones is at once apparent by (13).

Some measurements provided a very satisfactory experimental verification of the theory outlined above.

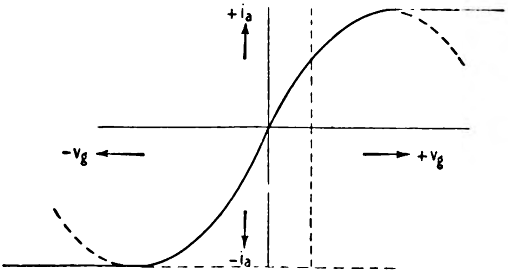

Fig. 2

The measurements were taken with relatively low frequency, thus avoiding obvious errors. That the results are, however, equally applicable to high-frequency circuits follows at once from the following theorem:

If a model is made of a high-frequency system consisting of linear inductances, linear capacities and non-linear resistances (e.g. triodes) and if the values of all the inductances (self and mutual) and capacities in the model are made n times these values in the original high-frequency system, but if the resistances (linear and non-linear) in the model are made equal to the resistances in the original circuit, the currents and potentials occurring in this model will be exactly equal in magnitude to the currents and potentials in the original high-frequency system but, considered as a function of the time, they will vary n times slower.[4]

Therefore, the natural periods of the model will be n times those of the original system, and the building up and decay of currents in the model will also occur n times slower.

This simple theorem at once follows from a dimensional consideration of the coefficients such as r, ωL, $1/\omega C$, $M\omega$, occurring in the differential equations, because, e.g.

$$\omega L = \frac{\omega}{n} \cdot nL, \text{ etc.}$$

[4] Incidentally, in this model system the stray capacities are reduced n times in magnitude. Therefore, in order to investigate the effect of a specified stray capacity in the original high-frequency circuit it is only necessary to insert at its place in the model circuit a capacity n times the original stray capacity.

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In fact, with a system having a natural frequency of 500 Hertz (cycles per second) and with critically adjusted regeneration it often took a minute for the free oscillation of the triode system to reach its final value, hence it was not easy to decide, before the external emf was applied, whether the regeneration was exactly critical or not. Therefore, an intermediate way was chosen and a model of a receiving set was made having a natural frequency of 15000 Hertz.

Four sets of readings were taken[5] with a standard triode (tungsten filament $V_f = 4.0$ volts, saturation current 11 milliamperes, anode potential varying between 100 and 200 volts, negative grid bias varying between -4.5 and -7.5 volts).

The applied emf $E \sin \omega_1 t$ was varied between 2.10^{-6} and 10^{-2} volts, the resulting alternating grid voltage V_{g1} with no regeneration varied between 100.10^{-6} and 0.5 volts, and the resulting alternating grid voltage b with critical regeneration varied between ca. 0.2 and 4 volts. The exponent s in the formula

$$b = (V^2_{g0} \cdot v_{g1})^s$$

which, according to the above theory, should be

$$s = 0.33$$

was found from the four sets of measurements to be

$$\begin{array}{r} s = 0.36 \\ 0.36 \\ 0.36 \\ 0.32 \\ \hline \text{mean: } s = 0.35 \end{array}$$

while the "grid space" V_{g0} calculated from these measurements was

$$\begin{array}{r} V_{g0} = 26 \text{ volts} \\ 25 \\ 20 \\ 15 \\ \hline \text{mean: } V_{g0} = 22 \text{ volts,} \end{array}$$

[5] The experiments and calibration of the necessary amplifiers were performed by Messrs. K. Posthumus and R. Veldhuyzen. Care was taken that the receiver did not react on the transmitter.

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as compared with the value

$$V_{g0} = 27 \text{ volts}$$

obtained directly from the characteristic.

The experimental value for the exponent s fits in well with the theoretical value; provided only those parts of the characteristic are considered for which no oscillation hysteresis occurs. The representation of the i_a, v_g characteristic, by a cubical parabola, with three constants S_1, S_2, and S_3 only, obviously cannot yield very accurate values for V_{g0} obtained with a voltage swing of not more than 0.35 volts.

Therefore, there is no doubt that for all practical purposes the theory given above fits in well with experiment.

Finally, we give some practical figures calculated from formula (12) with the following data:

$$\frac{\omega L}{r} = 40,$$

$$V_{g0} = 27 \text{ volts}.$$

Electromotive force E working in grid circuit	Resulting alternating grid voltage with no regeneration (v_{g1})	Resulting alternating grid voltage b, with critical regeneration	Amplification obtainable through critical regeneration (b/v_{g1})
10^{-6} Volts	0.04×10^{-3} Volts	0.31 Volts	7700
10^{-5} "	0.4×10^{-3} "	0.66 "	1600
10^{-4} "	4.0×10^{-3} "	1.4 "	360
10^{-3} "	0.04 "	3.1 "	77
10^{-2} "	0.4 "	6.6 "	16

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

RECEPTION EXPERIMENTS IN MOUNT ROYAL TUNNEL*

By

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Summary—*This paper deals with certain experiments carried out in the Mount Royal Tunnel of the Canadian National Railways at Montreal, Quebec, in order to determine how radio waves reach the receiving set. Preliminary experiments in 1926 indicated that the penetration of radio waves into the tunnel was a function of the frequency, short waves below 100 meters dying out within a few hundred feet of the mouth of the tunnel. More exact experiments and measurements were planned and carried out in 1928 to bring out the part played by the wires, cables, and rails leading into the tunnel. The tunnel mouths were blocked and the cables grounded, and the results so obtained indicated that the effect of cables and rails was also dependent on the frequency. It was also evident that more actual energy entered via the tunnel itself than was at first suspected. The effect of rails and cables is not a simple one but involves loop action, wave-antenna effects, and re-radiation.*

Curves are attached showing graphically the results obtained. A map of the area and an elevation of Mount Royal itself give general details of the geology of the region.

Introduction

EARLY in 1926 the problem of reception of radio transmission in the Mount Royal Tunnel in Montreal came up for consideration, and at the request of Dr. A. S. Eve arrangements were made to carry out a series of tests in order to determine the general nature and strength of radio waves under Mount Royal.

Information has been published from time to time in the *Technical Press* dealing with experiments of this nature both in England and America, but so far as is known, these experiments were confined to a judgment of signal strength by the ear and not to a measurement of the actual signal strength existing within the tunnel. It is a well known fact that the human ear is a very indifferent measuring instrument and it was therefore determined that in the proposed tests accurate measurements should be made.

* Dewey decimal classification: R113. Original manuscript received by the Institute, October 31, 1928. Presented before New York meeting of the Institute, December 5, 1928.

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Before considering the experimental work it might be well to spend a little time in a discussion of the tunnel itself. The attached section of Mount Royal, Fig. 1, shows the outline of the mountain as well as the location of the tunnel, and in addition an indication is given of the nature of the rock of which the mountain is composed. The tunnel is $3\frac{1}{2}$ miles long and extends from Dorchester Street to the C. P. R. tracks on the northwest side of the mountain. The diagram reproduced herewith has been taken from a report on the "Essexites of Mount Royal," by Bancroft and Howard, of McGill University. It will be noted

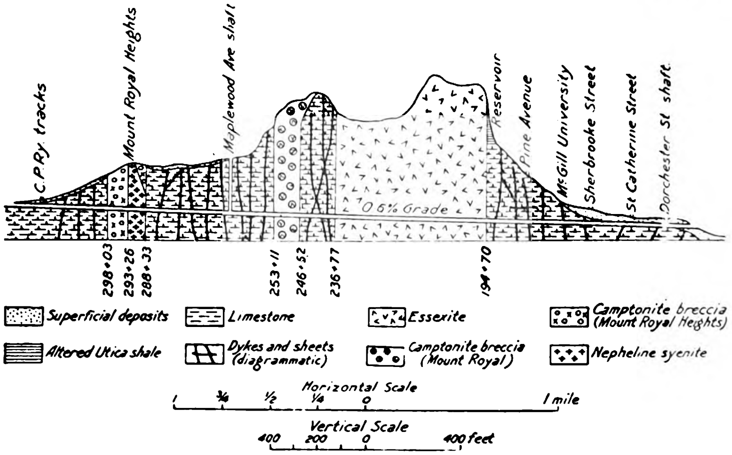

Fig. 1—Section of Mount Royal Showing Outline of Mountain and Location of Tunnel.

that the greater part of the mountain is composed of limestone, but the entire center section, that is, where the mountain is the highest, is composed of essexite. In addition to this there are one or two shafts of camptonite breccia in the northern part of the mountain. Attention is also drawn to the ventilation shaft at Maplewood Avenue.[1] This is the only ventilation shaft in the tunnel from Dorchester Street to the C. P. R. tracks. The tunnel is wide enough to carry a double track, and, in addition to the rails, the following conductors are carried through the tunnel from end to end:

(a) A lead covered conductor for 60-cycle, 12000-volt, 3-phase power.

[1] The tunnel also passes directly under the reservoir at Outremont not far from the ventilation shaft.

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(b) Two phosphor bronze trolley wires over each track. These trolley wires are suspended from a 7/8-in. diameter phosphor bronze messenger cable, the messenger cable itself being thoroughly grounded.

(c) A lighting circuit for the tunnel carrying 60-cycle single phase at 120 volts.

(d) The negative return wire for the ground connection.

(e) 20 or 30 telegraph conductors, lead covered and insulated for 240 volts.

(f) Between Dorchester Street and the grotto there is an automatic track circuit 1650 feet long. There are numer-

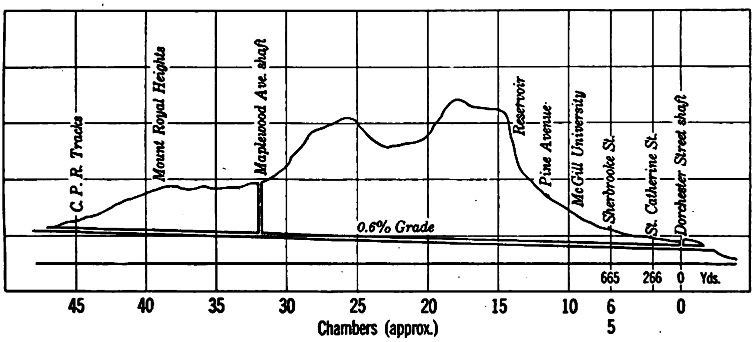

Fig. 2—Elevation, Mount Royal Tunnel, Montreal, Quebec, Canada.

ous signal and power wires required for this circuit, but the detail of the conductors is not definitely known.

(g) All four rails comprising the double track are thoroughly grounded with 250,000 circular mil stranded cable.

The tunnel is conveniently divided into chambers. By referring to Fig. 2 it will be noted that these chambers are not in all cases of uniform length. Chambers 5 and 6, for example, are both located at the same point in the tunnel, while Chambers 7 and 8 are very short. Again at the northwest end there are several chambers somewhat shorter than those in the central part. In general the chambers are 400 feet long. In plotting the curves for this report the base line has been taken as the subdivision in chambers, and an attempt has been made to take into consideration the essential differences mentioned above. This point, however, is not of very great importance in considering the results obtained.

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The map attached to this report shows in general the direction taken by the tunnel as it passes under Mount Royal. It will be noted that the actual direction of the tunnel is nearly east and west and the Montreal end is referred to as the "east portal" and the Mount Royal Heights outlet as the "west portal."

Preliminary Tests

In June, 1926, arrangements were made by McGill University, the Canadian National Railways, The Canadian Marconi Company, and The Royal Canadian Corps of Signals, for a series of reception tests in Mount Royal Tunnel. It was thought that tests should be made on different wavelengths in order to determine whether the frequency affected the penetration of the signals into the tunnel, and 40 meters, 411 meters and 1300 meters were finally chosen, these being representative wavelengths in their respective bands. Signals from station WIZ were to be used for the short-wave band, as this station was nearly always in operation and the signals were remarkably clear and constant in Montreal. The Canadian Marconi Company were good enough to arrange for a series of programs from their Montreal station, gramaphone records were used, and care was taken to maintain a constant output throughout each test. For the 1300-meter signals, the 500-watt station of The Royal Canadian Corps of Signals in Ottawa was employed. The tests began at midnight and an hour's transmission on each of the three wavelengths was required to make a series of tests from the Dorchester Street station to the west end of the tunnel. No attempt was made to measure the field strength in these tests, but a carefully calibrated audibility meter, connected as a shunt across the telephones in the output circuit, was employed for each of the three bands of wavelengths. It is appreciated that tests of this nature are not highly accurate, but they are at least an improvement over any attempt that might be made to judge the signals by the ear alone. The order in which these tests were made was as follows:

(a) From midnight until 1 A.M., Eastern Standard Time, 411-meter signals were received.

(b) Between 1 and 2 A.M. signals on 1300 meters were received from Ottawa.

(c) From 2 to 3 A.M. signals from WIZ were used.

For each of these wavelengths a standard car aerial was em-

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ployed. This aerial was of the box type and was mounted on top of the observation car used for the tests. For the short-wave reception the receiving set consisted of a detector and two steps of audio amplification. Marconi V.24 valves were used and regeneration was controlled by inductive coupling. A standard broadcast receiver of the neutrodyne type was used for the 411-meter wave. This was a 5-valve set operated without a power tube. An R. C. C. S. standard receiver was used for the long-wave band. This set had two stages of radio-frequency

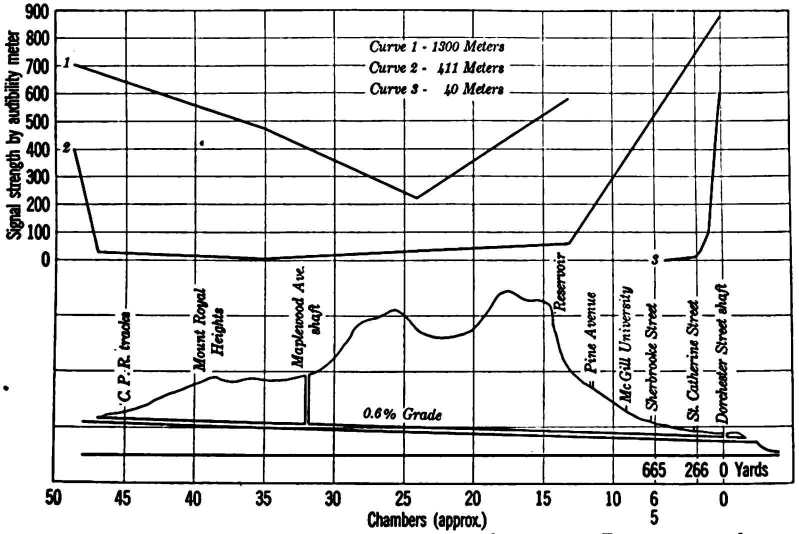

Fig. 3—Signal Strength, Measurements under Mount Royal Tunnel, Montreal, Quebec, Canada, June 16, 1926.

amplification and one stage of audio. On the radio-frequency side the coupling between valves was by means of the well known rejector circuit. In this receiver English "R" type valves were employed.

The procedure in taking all these tests was to start at the Dorchester Street terminus and to proceed by definite steps through the tunnel. In the case of the broadcast and long-wave bands, these steps were each approximately 10 chambers long. In the case of the short-wave signals, very much shorter steps had to be employed as the signals died out within a few hundred feet of the entrance to the tunnel.

The results of these experiments can best be followed by a study of Fig. 3. In this diagram the profile of Mount Royal

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itself has been laid down to the same base as is used for the plotting of the results. In this way it was possible to identify a given signal strength with the depth of rock existing over the tunnel at that point. The most noteworthy result was the rapid attenuation on the short-wave band. It will be observed that the signals were reduced to zero approximately 1500 feet from the mouth of the tunnel, that is, before Chamber 5 was reached. This result was so unexpected that repeated checks were made, the signal being traced step by step into the tunnel. The net result, however, in each case was as shown by Curve 3. It will be observed

TABLE I

42.5-Meter Tests—April 14, 1928

Position in Tunnel	Control Panel Reading in Current Ratios	Per Cent Signal Strength (5)	Remarks
300 ft. N of Chamber 45	13	100	
Chamber 45	10	76.8	
100 ft. S of Chamber 45	5	38.4	
150 ft. S of " 45	2.5	19.2	
200 ft. S of " 45	2.0	15.4	
300 ft. S of " 45	1.6	12.3	
Chamber 44	6.2	47.7	
100 ft. S of Chamber 44	2.0	15.4	
Chamber 43	1.0	7.7	
" 42	0.39	3.0	
" 41	0	0	

TABLE II

42.5-Meter Tests—April 15, 1928

		(8)	
300 ft. N of Chamber 45	3.1	100	
100 ft. S of " 45	1.25	40.4	
100 ft. S of " 45	1.6	51.5	
200 ft. S of " 45	0	0	

that there were only 48 feet of rock over the tunnel at the point where the short-wave signals disappeared. These tests were only made at the east or Dorchester Street end of the tunnel.

The 411-meter signals were audible throughout the entire length of the tunnel, but were very weak from Chamber 11 to Chamber 44. At Chamber 11 the strength was reduced to one twelfth of its maximum strength outside the tunnel. At this point the rock was approximately 275 feet deep. It should also be noticed that at Chamber 33, where the broadcast signals are weakest, the depth of rock is only 210 feet.

The 1300-meter signals were also audible throughout the tunnel, but were very much stronger than the broadcast signals although the two stations were of approximately the same power output. It should be noted, however, that the station from which the

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1300-meter signals were received was about 100 miles from the tunnel, while the broadcast station was only about 3 miles away. In this case the signals reached their lowest value at Chamber 22, which is approximately the center of the tunnel.

From a consideration of these curves it would appear that the strength of signal at any point within the tunnel was a very definite function of the frequency employed. It still remained to be determined, hcwever, whether these signals came through the open tunnel or along the conductors and rails running through the tunnel. Considerable discussion arose over this

TABLE III
55-Meter Tests—April 14, 1928

Position in Tunnel	Meter Reading	Per Cent Signal Strength (3)	Remarks
300 ft. N of Chamber 45	400	100	
Chamber 45	200	50	
100 ft. S of Chamber 45	10	2.5	Fading
150 ft. S of " 45	20	5.0	
200 ft. S of " 45	15	3.75	
300 ft. S of " 45	9	2.25	
Chamber 44	15	3.75	
100 ft. S of Chamber 44	12	3.0	
Chamber 43	10	2.5	
" 42	5	1.25	
" 41	1	0.25	

TABLE IV
55-Meter Tests—April 15, 1928

		(6)	
100 ft. S of Chamber 45	100	100	
200 ft. S of " 45	15	15	
300 ft. S of " 45	50	50	
Chamber 44	40	40	
" 43	30	30	
" 42	8	8	
" 41	0	0	

point and two general theories were evolved as a result of the work done. The first theory was that the greater part of the energy reached any given point within the tunnel by passing through the rock. It was appreciated that a certain percentage must come along the wires and perhaps through the tunnel itself. In support of this theory the following figures are of interest. From Zenneck's treatise on "Radio Telegraphy," page 248, the following figures with regard to conductivity are obtained,—

Conductivity of sea water, $\delta = 1$ to 5×10^{-11} E. M. Units
Dryish earth, $\delta = 10^{-14}$ E. M. Units
Ratio = 1 to 1000

From the same authority we learn that one meter of sea water reduces the strength of signals by 10 per cent. Therefore, it is

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assumed that it would require 1000 meters of rock to produce the same reduction in strength.

The other theory advanced considered that the greater part of the energy entered the tunnel via the tunnel mouth, although a proportion undoubtedly came along cables and rails. The curves of Fig. 3 afford the best support for this theory. From these curves it will be noted that the short wave dies out within a few feet of the mouth of the tunnel although there exists only a thin layer of rock at this place. Again, it will be noted that the 411-meter signal continues to diminish in strength as we go from the Dorchester Street shaft towards the west end of the tunnel and at Chamber 33 the lowest signal was received, although the rock is considerably shallower at this point than it is in the center of the tunnel. From Chamber 33 on, the signal increases in strength, but this might be accounted for by energy coming in via the west portal.

The 1300-meter signal is the only one that seems to follow the rock penetration theory; here the weakest signal was received under the center of the mountain, that is, under the greatest depth of rock.

On account of shortage of time it was not possible to carry out further experiments during 1926, in order to determine which theory was the more nearly correct.

During the summer of 1927 Dr. Eve was in a position to carry out further tests in the Caribou Mine of the American Mining and Prospecting Company at Caribou, Colorado.[2] The following extract from this report will indicate quite clearly the tests made and the results obtained—

> The experiments participated in by Dr. Eve were conducted with a superheterodyne set with nine electron tubes in the Caribou mine of the American Mining and Prospecting Company, at Caribou, Colorado. The first test was held at a depth of 220 feet, where, by means of a loop, a strong and clear reception was obtained of a musical concert given at Denver, 50 miles distant. The evidence pointed strongly to the conclusion that this clear reception was due to the penetration by the radio waves of the solid rock strata, although there was a remote possibility that the reception was obtained through shafts and crosscuts, toward which, however, the loop did not point. The nearest metal conductors, iron rails, were 66 feet away.

[2] A resumé of these tests appeared in the *Jour.* A.I.E.E., November, 1927. See also Technical Papers 434, pages 37-40, Dept. of Commerce, Bureau of Mines, Washington, D. C., U. S. A.

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The next series of experiments was conducted at a depth of 550 feet, when "mushy" reception was obtained from Denver. This type of reception was, however, as good as could be obtained above ground at the time of making the test, the night being unfavorable for general radio reception. This series of tests was conducted at the end of a cross-cut reached with many turns, and 200 feet from the main shaft. A pipe came down the shaft and followed the tunnel up to 80 feet from the point of observation.

In previous experiments conducted by the Bureau of Mines at its Experimental Mine near Pittsburgh, Pa., it was at first concluded that radiation and induction would penetrate rock for considerable depths. Subsequent investigations have shown that in every case the transference of radiation was by some conductors in the mine, electric wires, pipes or rails, all of which abound in modern mines.

It is felt that further investigations should include a comparison of the penetration of radio waves greater and less than a wavelength.

It is pointed out that in making these tests no attempt was made to assist the ear in determining the relative signal strengths. It is a known fact that the unaided human ear is a relatively poor test instrument. The following facts in this connection should be kept in mind when comparing the results of the two

TABLE V
411-Meter Tests—April 14, 1928.

Position in Tunnel	Meter Reading	Per Cent Signal Strength (1)	Control Panel Reading in Current Ratios	Per Cent Signal Strength (2)
Dorchester Station	42	100	12.5	100
Chamber 5	11	36.2	6.2	49.6
" 10	7	16.7	2.0	16.0
" 15	5	11.9	2.0	16.0
" 20	2	4.75	1.25	10.0
" 30	4.5	10.7	2.0	16.0
" 40	22	52.4	12.5	100
" 45	18	42.9	0.39	3.12
300 ft. N of Chamber 45	45	107.0	10.0	80

TABLE VI
411-Meter Tests—April 17, 1928.

		(10		(11)
Chamber 1	3000	100	13.9	96.5
" 5	500	18.3	13.0	90.3
" 10	350	11.65	13.0	90.3
" 15	80	2.67	5.0	34.7
" 20	60	2.0	4.0	27.8
" 25	300	10.0	6.2	43.0
" 29	500	16.7	13.0	90.3
" 30	800	26.7	6.2	43.0
" 31	600	20.0	0.8	5.55
" 32	350	11.65	0	0
" 35	550	18.3	3.2	22.2
" 40	400	13.3	7.9	54.8
" 44	500	16.7	14.4	100

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separate series of tests. Fig. 4 is a curve taken by means of a Shaw fading recorder and represents the actual carrier waves being received at the receiving station. Particular attention should be paid to the curve representing the carrier wave from WEAF on the night of April 10th, 1928. The signal in the loud-speaker due to the carrier wave, and represented by the height from the zero line 0, to the dotted line *A*, was a good average

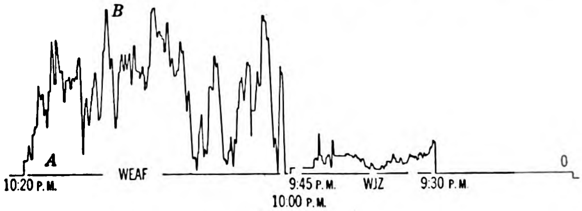

Fig. 4

signal, loud enough to be heard throughout an ordinary room. The signal equivalent to the peaks of the wave, while much stronger, did not sound loud enough to bear the proportion indicated by the relative heights of the corresponding lines. In fact, it is doubtful if the human ear would notice that the signal had changed by more than 50 per cent between the point *A* and the point *B*. Again referring to the tables of results in Appendix 1, attention is called to column 10, Table VI. In Chamber 1 the strength

TABLE VII
1400-METER TESTS, APRIL 14, 1928.

Position in Tunnel	Meter Reading	Per Cent Signal Strength (4)	Remarks
Dorchester Station	1000	100	
Chamber 11	300	30	
" 19	200	20	
" 31	400	40	
" 39	800	80	

TABLE VIII
1400-METER TESTS, APRIL 15, 1928.

		(7)	
Dorchester Station	1000	100	
Chamber 1	300	30	
" 11	150	15	
" 19	200	20	
" 31	400	40	
" 39	1000	100	
" 45	1500	150	
300 ft. N of Chamber 45	5000	500	

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of signal is represented by a meter reading of 3000 while at Chamber 20 this had dropped to a reading of 60. To the human ear this did not appear to be a change of more than 100 per cent, when estimating the relative strengths of the signals coming from the loudspeaker.

In judging the results of the Colorado tests from the point of view of the theories already advanced, it would appear that they seem to strengthen the impression that the energy reached the receiver by passing through the ground or rock. It must be remembered, however, that no attempt was made to measure the strength of the carrier wave and that the ear alone has been shown to be a very poor measuring instrument. The differences have already been enumerated. It would, therefore, be inter-

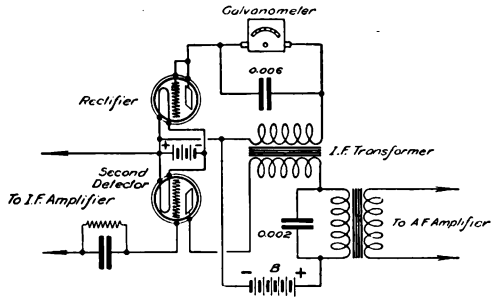

Fig. 5

esting to compare the results obtained in these two tests, insofar as the relative thickness of the covering is concerned. According to the Colorado figures radio waves should penetrate earth and rock to a depth of at least 550 feet. In the first tunnel experiments it was found that 48 feet of rock were sufficient to reduce the 40-meter wave to zero. In the case of the 411-meter wave, which was approximately the wavelength used in Colorado, the signal was reduced to one twelfth of its maximum strength by 275 feet of rock. In addition to this, the curves of Fig. 3 show that the strength of the 411-meter wave was of a very low order from Chamber 11 to almost the west portal of the tunnel although from Chamber 28 onward the thickness of rock is steadily decreasing. It is hardly fair to compare the 1300-meter results under Mount Royal with the Colorado tests as it is evident from the work already carried out that frequency has a very important bearing on the rate of attenuation in the tunnel.

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Tests of April, 1928

In view of the conflicting results obtained in the preliminary tests it was decided to carry out a series of experiments, during the spring of 1928, in the Mount Royal Tunnel and to endeavor to settle definitely the various problems which arose as a result of the first work undertaken. Assuming a given strength of signal at any point within the tunnel three problems had to be solved—

(1) The percentage of the signal arriving by conduction through the rock.

(2) The percentage of the signal being conducted along rails and cables leading into the tunnel.

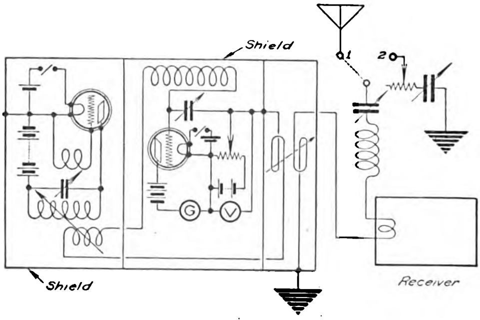

Fig. 6

(3) The percentage of signal entering the tunnel mouth and being conducted as in the case of waves passing over the surface of the ground.

From the description already given of the tunnel it will be appreciated that the separation of these three effects presented a difficult problem. It was not possible to take out the cables completely and while a section of the rail could have been removed, it is doubtful if any small section would, in itself, show an appreciable effect in view of the comparatively small dimensions of the tunnel and the large number of electrical conductors passing through it. After careful consideration it was decided to carry out the following three series of tests—

(1) With all rails, wires, and power conductors as for normal operation of the tunnel, tests were to be made on four wavelengths. These wavelengths were to be located in the following bands—

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(a) Short wavelength band below 100 meters.
(b) Broadcast band.
(c) Medium wave band between 1000 and 2000 meters.
(d) Long-wave band above 10,000 meters.

(2) On the second night the same series of tests were to be run, but steel coaches were to be placed in each end of the tunnel so as to block both portals as far as possible.

(3) On the third night the same series of tests were to be made with the tunnel blocked and with all power, light,

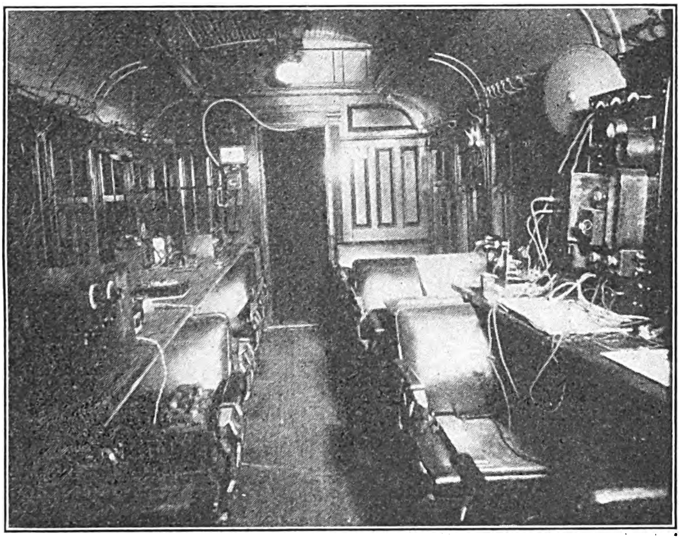

Fig. 7

and signal cables thoroughly grounded. In addition to this the overhead trolley wires were to be grounded at two points, one at either end of the tunnel.

It is appreciated that these tests would not constitute a complete solution of the problems, but it was hoped that they would, at least, indicate the general nature of the transmission. A wooden coach was used to carry the equipment, the standard pullman car aerial being used for the medium and long-wave results, but a special short-wave antenna was mounted inside the car for the short-wave signals.

In assembling the equipment for this work it was kept in mind that in each case the actual strength of the carrier wave was to be measured. This was not possible in the case of the short-

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wave receiver, as a field strength measuring set was not available for this band. In place of this a thermionic voltmeter was used to measure the actual voltage developed across an inductance coil in the output circuit. In addition to this method of measurement, an audibility meter was available so that the receiving set could be thrown from one to the other and check results rapidly obtained. For the broadcast reception a superheterodyne set with a loop receiver was employed, the strength of carrier wave being measured by the method described by Dr. G. W.

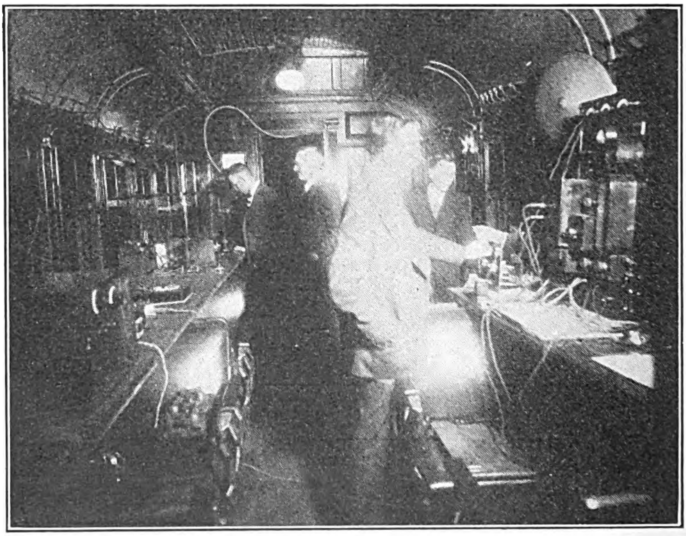

Fig. 8

Pickard.[3] This scheme consists essentially of introducing an additional intermediate frequency transformer into the second detector circuit of the superheterodyne. The output from this transformer is rectified and measured on a sensitive galvanometer. Fig. 5 shows the general principle of the method.

In addition to these two methods of measurement for the short and broadcast waves, a standard Western Electrical control panel was set up by the Canadian National Railways Radio Department and used with standard receivers to check the strength of the received signals. These results were measured in transmission units, and transferred to current ratios, for plotting in the form of curves.

[3] G. W. Pickard, *Radio Broadcast*, August, 1926.

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The medium-wave signals were measured by using a thermionic voltmeter across an inductance in the detector circuit, the current in this inductance being proportional to the strength of the carrier wave. In this case the audibility meter was also available for purposes of checking.

For the long wave the Canadian Marconi Company supplied a signal strength measuring set similar to the type used by Captain Round during his expedition to Australia in 1922 and 1923. Fig. 6 is a schematic diagram of the equipment used.[4]

Figs. 7 and 8 attached will show the set-up used in the tunnel tests. These figures are from photographs taken inside the car while the measurements were being made. Referring to Fig. 7, the medium wave set may be observed in the foreground on the

TABLE IX

1400-Meter Tests, April 17, 1928.

Position in Tunnel	Meter Reading	Per Cent Signal Strength (12)	Per Cent Signal Strength (12a)	Remarks
Dorchester Station	1000		100	
Chamber 1	500	100	50	
" 5	800	160	80	
" 10	800	160	80	
" 15	150	30	15	
" 20	800	160	80	
" 25	1000	200	100	
" 29	200	40	20	
" 30	150	30	15	
" 31	200	40	20	
" 32	800	160	80	
" 35	600	120	60	
" 40	500	100	50	
" 44	2000	400	200	

left hand side of the photograph. Immediately behind it is the loop and broadcast receiver while on the right of the photograph may be seen the Western Electric control panel and one of the short-wave receivers. On account of its size the Marconi field strength set had to be mounted in the after compartment of the car.

In order to minimize the effects of fading and changes in power output at the transmitting station, it was decided to utilize, as far as possible, transmitting stations over which control could be exercised. Short and medium-wave signals were available from the Ottawa stations of The Royal Canadian Corps of Signals. As previously stated these stations were located about 100 miles from Montreal. Previous tests had shown that 55 meters was a

[4] A complete description of the method of operation will be found in the *Jour.* I.E.E., October, 1925.

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very satisfactory wave for Montreal and this was therefore decided upon for these tests. The medium-wave transmission was on a wavelength of 1400 meters. In both of these cases the power output to the antenna was maintained at a constant value of 750 watts both for the short wave and for the medium wave signals. The broadcasting station of the Northern Electric Company in Montreal was secured for broadcast transmission on 411 meters. Arrangements were made to hold the output and modulation of this station very constant. The modulation was reduced

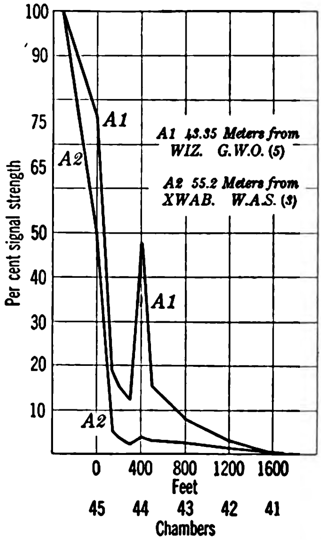

Fig. 9—Comparison of Signal Strengths. Short-Wave Tests, April 14, 1928; Tunnel Open.

from that normally employed and was approximately 15 per cent during the actual tests.

As there was no long-wave station in Canada that could be used for the 10,000-meter results, it was decided that one of the Radio Corporation stations on Long Island would be copied. The stations actually used for these tests were WCI on 16,700 meters, WSS on 16,120 meters, and WQK on 16,465 meters. The tabulated results from these stations will be found in Appendix 1, Tables X, XI, XII, and XIII.

The program followed for each of the three nights of the test was as follows; from midnight to 1 A.M. broadcast reception was measured. From 1:15 to 2:15 A.M. short-wave signals were

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received. From 2:30 to 3:30 both medium-wave signals and long-wave signals were checked as it was possible to carry out both of these measurements without interfering one with the other

In order to check possible errors due to natural fading, observers were stationed in and around Montreal for the purpose of recording the approximate strength of the various stations throughout the time of the test in the tunnel. Fortunately, it was found that there was very little natural fading during the three nights. As the medium and long-wave signals are not so

TABLE X
LONG-WAVE TESTS, APRIL 14, 1928.

Position in Tunnel	Reading in Microvolts	Per Cent Signal Strength (13)	Station and Wavelength	Remarks
Dorchester Station	233	100	WSS	Power on
Chamber 39	160	68.7	16,120	Moving North
200 ft. N of Chamber 45	1530	656	meters	
		(14)		
Dorchester Station				
Chamber 11	600	257	WSS	Power off
" 19	435	186	16,120	Moving North
" 31	870	373	meters	
" 39	870	373		
200 ft. N of Chamber 45	1530	656		
		(15)		
Dorchester Station	1100	472	WSS	Power on
Chamber 11	600	258	16,120	Moving South
" 19	260	111.5	meters	
" 31	2050	880		
		(16)		
Dorchester Station	600	258	WSS	Power off
Chamber 11	1300	558	16,120	Moving South
" 19	1300	558	meters	
" 31	2500	1071		

much affected by fading as are the broadcast and short-wave bands, no check record was kept of the signal strengths either for the 1400-meter or the 16,000-meter stations.

The following brief description of the method of preparing the tunnel for the tests on the second and third nights may be of interest. The steel cars were placed just inside the mouth of the tunnel where the tube has its smallest diameter. It was estimated that approximately 80 per cent of the area of the tunnel mouth was blocked by the two steel cars. The power cables were treated as follows,

(a) The 110-volt a.c. lighting circuit was opened at both ends by means of switches, but was not grounded.

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(b) The 2400-volt d.c. feeder was opened at the sub-station. From the sub-station to the tunnel mouth there is a feeder approximately 1000 yards long. This feeder was not opened at the tunnel. The 2400-volt trolley wire terminates in the Dorchester Street station at the East end of the tunnel. This trolley wire, together with the supporting cable, was grounded to the bonding cable at Chambers 5 and 43. These points were not quite at the two ends of the tunnel, but it was convenient to make a thoroughly satisfactory ground at these two points.

(c) The 12,000-volt 3-phase power cable which supplies power to the sub-station and to the village of Mount Royal enters at the Dorchester Street Station and runs through the tunnel. It is carried in a duct between the two halves of the tunnel. This cable is covered with a lead sheath and the lead sheath was grounded at several points, but the power was not cut off during any part of the tests.

(d) All telegraph and signal wires are carried through the tunnel in lead covered cable. These wires were not opened at any time during the test, but the lead sheath was at all times thoroughly grounded in several places.

Complete tables of the results of the April tests are shown. As already pointed out more than one system of measurement was used. In view of this difference in methods all data have been reduced to the basis of "Per Cent Signal Strength," taking as the unit the strength of signals received in the Dorchester station, with each different system of measurement; that is, the signal received at this point was considered as a 100 per cent signal. The columns marked "Per Cent Signal Strength" are numbered from 1 to 21 inclusive, and the same numbers will be found on the curve sheets attached. Tables I to IV, inclusive, cover the short-wave tests, Tables V and VI the broadcast tests, Tables VII to IX, inclusive, are for the 1400-meter signals, and Tables X to XIII, inclusive, are the long-wave results. Thirteen curve sheets, Figs. 9 to 21, have been drawn based on these tables. In these curves,

A represents short-wave results,
B broadcast or 411 meters,
C 1400 meters,
D 17000-meter readings.

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TABLE XI

Long-Wave Tests, April 15, 1928.

Position in Tunnel	Reading in Microvolts	Per Cent Signal Strength (17)	Station and Wavelength	Remarks
Dorchester Station	325	100	WCI 16,700 meters	Power on Moving South
Chamber 1	163	50		
" 11	455	140		
" 19	130	40		
" 31	490	151		
" 39	130	40		
" 45	360	111		
300 ft. N of Chamber 45	1170	360		
		(18)		
Dorchester Station	500	154	WCI 16,700 meters	Power off Moving South
Chamber 1	910	280		
" 11	1170	360		
" 19	780	240		
" 31	780	240		
" 39	780	240		
" 45	1050	323		
300 ft. N of Chamber 45	1170	360		

TABLE XII

Long-Wave Tests, April 16, 1928.

		(19)		
Chamber 1	65	100	WCI 16,700 meters	Power off Going North
" 11	160	246		
" 19	130	200		
" 31	520	800		
" 39	65	100		
" 45	110	169.5		

TABLE XIII

Long-Wave Tests, April 17, 1928.

Position in Tunnel	Readings in Microvolts	Per Cent Signal Strength (20)	Station and Wavelength	Remarks
Chamber 1	130	100	WCI 16,700 meters	Tunnel blocked cables grounded.
" 5	32	24.6		
" 10	0	0		
" 15	1170	900		
" 20	780	600		
" 25	1800	1382		
" 29	1800	1382		
" 30	2000	1540		
" 31	2000	1540		
" 32	1800	1382		
" 35	15[illegible]0	424		
" 40	1170	900		
" 44	910	700		
		(21)		
Chamber 1	65	50	WQK 16,465 meters	Tunnel blocked cables grounded.
" 5	30	23.1		
" 10	65	50		
" 15	540	415		
" 20	230	177		
" 25	540	415		
" 29	780	600		
" 30	980	754		
" 31	980	754		
" 32	780	600		
" 35	260	200		
" 40	260	200		
" 44	160	123		

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Discussion of Results of 1928 Tests

The results obtained in the above mentioned tests can most easily be studied by means of the curve sheets mentioned above.

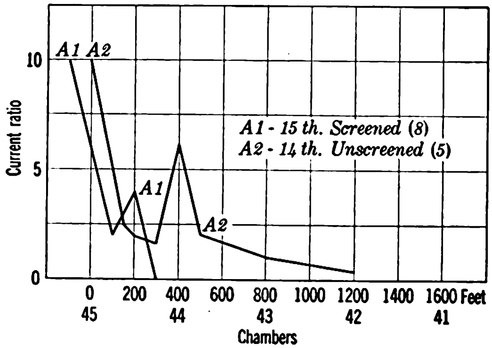

Fig. 10—Comparison of Signal Strengths.
42.5-meter Tests, April 14-15, 1928.

Fig. 9 is a preliminary comparison of the short-wave results on 43 and 55 meters. Curve A1 was taken by means of the control panel and Curve A2 by the vacuum-tube voltmeter. Atten-

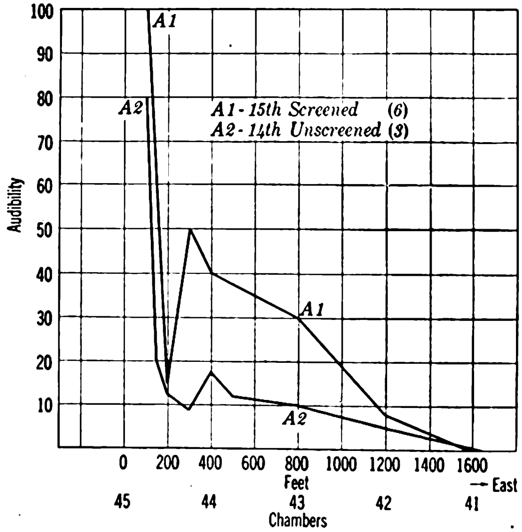

Fig. 11—Comparison of Signal Strengths.
55-meter Tests, April 14-15, 1928.

tion is called to the peculiar jump which occurred in both curves at Chamber 44, that is, 400 ft. inside the west portal. This station was immediately under the C.P.R. tracks. It is rather

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interesting to note that both waves act similarly and that both disappear at approximately the same point, although the signals were from stations located at widely distant points and varying greatly in power output. It might, therefore, be assumed that the attenuation on all waves below some certain frequency was approximately the same.

Figs. 10 and 11 have been drawn to show the effect on the short wave of screening the tunnel mouth by means of steel cars. In Fig. 10, it would appear from the 43-meter results that the screening had produced a decided decrease in strength. Fig. 11

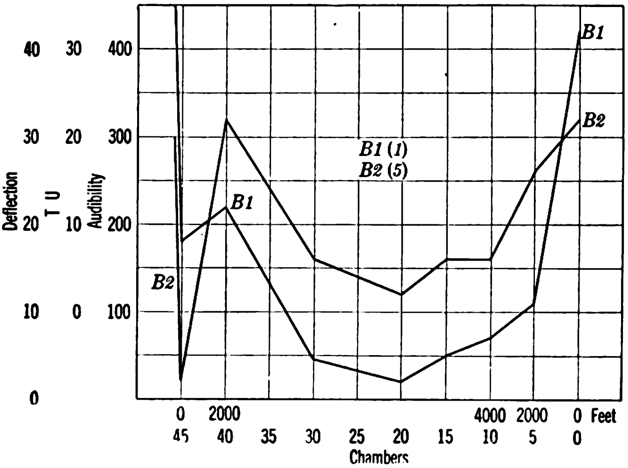

Fig. 12—Comparison of Signal Strengths.
411-meter Tests, April 14, 1928.

would seem to contradict this statement. There is at present no satisfactory explanation for this difference unless it may be accounted for by experimental error and the fact that the decrease in strength on the short wave is very rapid. The records at the transmitting station indicate that there was no appreciable change in power output during these tests.

Fig. 12, has been drawn for the broadcast band to show that while the results of different methods of measurement may vary slightly numerically, the nature of the effect is the same in all cases. Curve B1 was taken using the loop and superheterodyne and is a measurement of the actual carrier wave strength. Curve B2 was taken using the Western Electric control panel and is, therefore, more nearly a measure of the audio output of the receiving set. From Figs. 9 to 12 it may, therefore, be assumed

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that the general comparison of results of different methods of measurements is justifiable.

Fig. 13 is plotted for the broadcast band and brings out markedly the loss in signal strength due to blocking the tunnel and grounding the cables. The effects plotted here are the combined effects of the two actions. This was unavoidable due to the fact that it was impossible to make a 411-meter test on the 15th when only the blocking of the tunnel was being carried out. Attention is called to the sudden jump in Curve B2 at Chamber 30. This will be commented on later when dealing with the effect

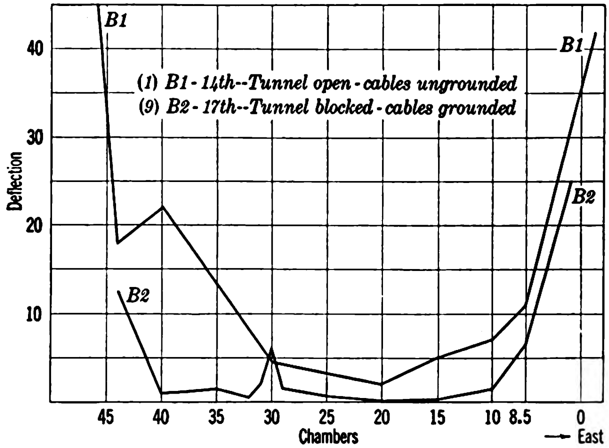

Fig. 13—Comparison of Signal Strengths. 411-meter Tests, April 14 and 17, 1928.

of the ventilation shaft at Maplewood Avenue. Figs. 13 and 14 are essentially the same, the only difference being that in Fig. 13 the actual deflection was taken as the ordinate, while in Fig. 14 the ordinate is the per cent signal strength. Curve B2 is the same in Figs. 13 and 14, but in Fig. 14 it is plotted from the audibility meter reading instead of the thermionic voltmeter reading.

An examination of the profile of the tunnel will show that the Maplewood Avenue ventilation shaft enters at about Chamber 30. This is the point at which the sudden jump occurs in the curves representing the results for the blocking of the tunnel and the grounding of the cables. It would, therefore, appear that sufficient signal comes in via this shaft to boost the curve very appreciably. This is a very interesting point as the shaft is only 30 ft. square and almost 300 ft. long. This fact considerably

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strengthens the theory that the signals enter the tunnel via the mouth rather than through the rock. The same effect will be noticed in the case of the long-wave results in Figs. 16, 17, and 18. In the case of the 1400-meter signals, Fig. 15, Curve C3, there would appear to be some additional action effecting the

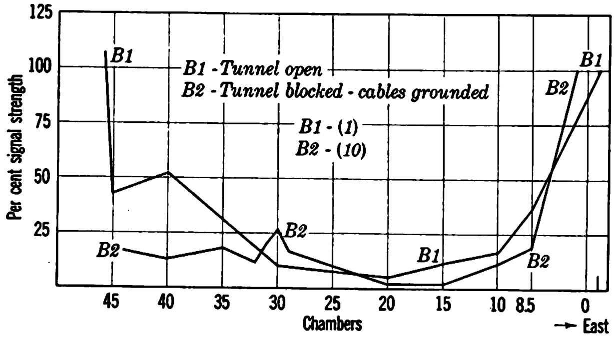

Fig. 14—Comparison of Signal Strengths. 411-meter Tests, April 14-17, 1928.

results. It is assumed that this is due to the loop action of the grounded cables and will be referred to in more detail when considering the long-wave results. It should be observed that the effect of the Maplewood shaft is only apparent when both screening and grounding are being employed. This is to be ex-

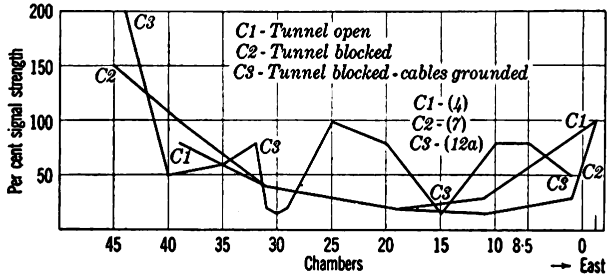

Fig. 15—Comparison of Signal Strengths. 1400-meter Tests, April 14, 15, 17, 1928.

pected as under ordinary conditions the signals enter through the main tunnel more readily than through the vertical shaft.

While studying the results of the tests, a suggestion was made that pipes, ladders, or steel work in the ventilation shaft might account for the conduction of signals down the shaft. At our request, a very careful search was made by Mr. Price, the elec-

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trical superintendent of the Mount Royal Tunnel, and he reports that there are no pipes, wires, hoist guides, or cables in the shaft, and in fact, no metallic substances of any nature of sufficient size to be worth considering.

Fig. 17 has been drawn from the long-wave results and demonstrates that the effect of the air shaft is similar in the case of the

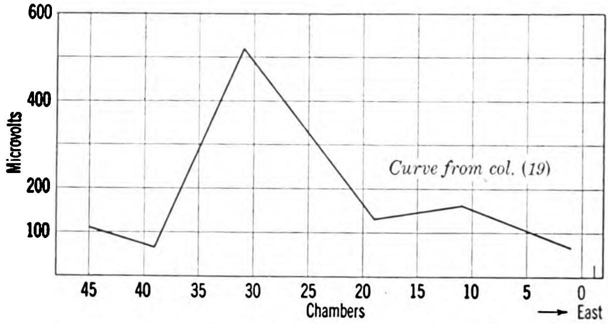

Fig. 16—WCI—17,000-meter Tests, April 16, 1928.

two long-wave stations used in the tests, although they have different output characteristics and different wavelengths. It would appear from the curves of Fig. 17 that the system of signal strength measurement used by the Marconi Company for the long-wave results was very reliable.

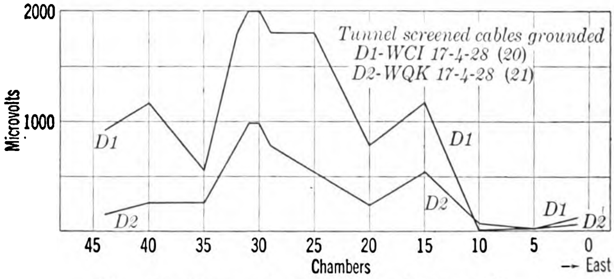

Fig. 17—Comparison of Signal Strengths.
17,000-meter Tests, April 17, 1928.

The relative effects of screening the tunnel and grounding the cables for long waves can readily be studied from Fig. 18. With the exception of Curve D3, the results are very consistent. In this curve the jump due to the air shaft is very marked. Attention is also called to the second peak at Chamber 15. This is a further evidence of the loop action to be discussed later.

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In order to obtain a comparison on one sheet of the relative effects for all frequencies, Figs. 19, 20, and 21 have been drawn. Fig. 19 compares the results with the tunnel open, Fig. 20 shows the signal strengths with only the blocking in place in the tunnel mouth, while Fig. 21 shows the total effect of blocking and screening being carried out together.

It has already been mentioned that the 2400-volt power feeder was alive during part of the test. As a matter of experiment, the power was cut off on several occasions and it was discovered

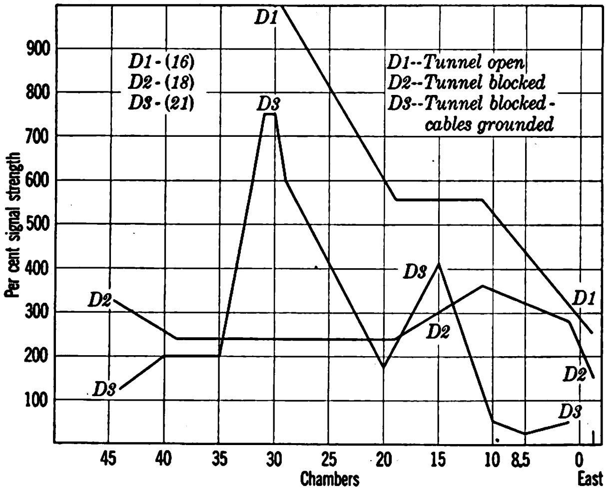

Fig. 18—Comparison of Signal Strengths. 17,000-meter Tests, April 14, 15, 17, 1928.

that there was a marked difference in signal strength with the power off, on certain frequencies. On the short waves there was practically no effect between power on and power off. On the broadcast wave the ratio of signal to noise was much better when the power was off. This, of course, was to be expected. On the 1400-meter and 17000-meter waves the signal strength actually seemed to be greater with the power off. This may be observed by comparing columns 13 and 14, 15 and 16, and 17 and 18, of Tables X and XI.

Referring again to Figs. 19, 20, and 21, with the tunnel open the per cent signal seems to be a direct function of the frequency, with the weakest signal in the case of each wavelength occurring at the center or about Chamber 20. The profile of the tunnel

shows that this chamber is just about midway between the two highest points of the mountain. With the tunnel blocked the same relative effects occur but the long-wave signals reach their minimum at a different point from the 1400-meter signals. Fig. 21 has been drawn for both blocking and grounding On this sheet the 55-meter signals have been omitted for reasons already stated. The effect of the air shaft is clearly indicated. There is, however, evidence of a further effect. It will be observed that in both the 1400-meter and the 17000-meter curves, peaks occur at regular

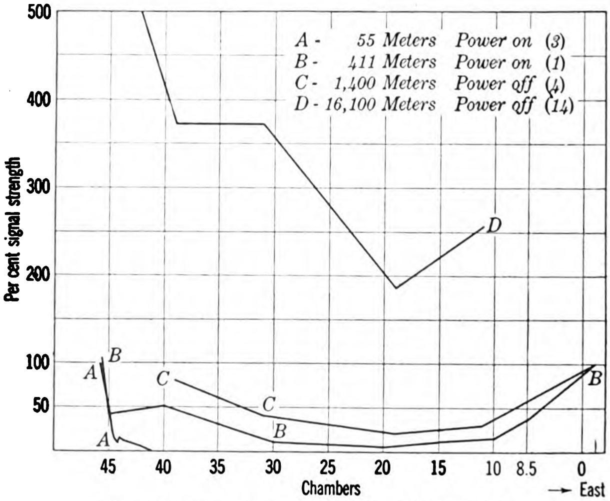

Fig. 19—Comparison of Signal Strengths, April 14, 1928. Tunnel Open.

intervals. These are not observable when the cables and trolley wires are ungrounded. It will be remembered that the grounding was carried out at Chambers 5 and 43, or approximately at the two ends of the tunnel. It would appear as if the grounding of the overhead trolley and cables had turned the system into a huge loop. It is rather difficult to calculate the characteristics of this loop on account of the nature of the construction. The messenger cable supporting the trolley wires is itself grounded. This brings the actual ground connection into close proximity with the overhead wires. There is, however, a 250,000 circular mil cable used to ground the tracks and this would, of course, constitute the other side of the loop. Whatever the characteristics of such a loop, it is quite conceivable that in a loop of this length,

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(approximately 3 miles) nodes and loops would occur for the different frequencies. From the curves it would appear that there were about twice as many loops for the 1400-meter signals as for the 17000-meter signals.

In connection with the long-wave signals it will be immediately evident that there is a very great difference between the results in the Dorchester station and those in the vicinity of the west portal. This would hardly be expected when it is considered that the Dorchester station is practically in the open at the east end of the tunnel, whereas the signals pass completely over or through the mountain to reach the vicinity of the west

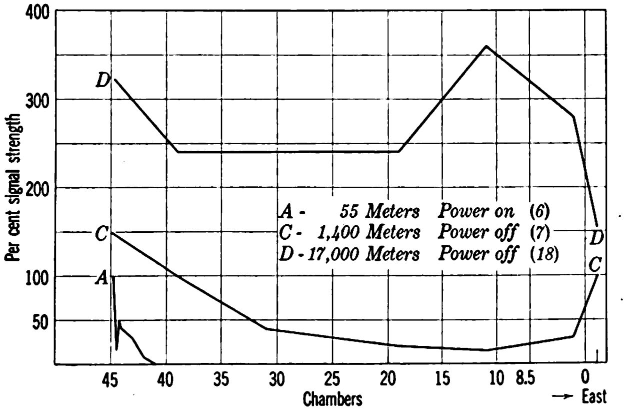

Fig. 20—Comparison of Signal Strengths, April 15, 1928. Tunnel Blocked.

portal. It is suggested that the action indicated in Tables X to XIII, and in the corresponding curve sheets, Figs. 18 to 21, is due to a wave-antenna effect. In other words, the overhead trolley wires which are open at the east end are acting as a Beveridge antenna with the east or open end pointing in the general direction of the transmitting station. If this assumption is correct the maximum effect will then occur when the receiving set is located at the west end of the feeder. It will, therefore, be seen that the results obtained fit in with the suggested theory. This effect would not be so noticeable on the 1400-meter signals since the transmitting station is located in a westerly direction from Montreal.

In this discussion little has been said by way of explanation of the peculiar jumps occurring in the curves in the vicinity of

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Chambers 40 to 45. It is quite likely that these effects are in some measure due to reradiation from the numerous telegraph, telephone, and power wires which run along the C.P.R. tracks above this section of the tunnel. This would not explain all the effects observed, but must be given some consideration as the various lines extend for considerable distances across country.

Attention is called to a further fact observed during the broadcast reception. It will be remembered that a loop was used for these signals and it was found that upon entering the tunnel the loop had to be turned parallel with the axis of the tunnel in

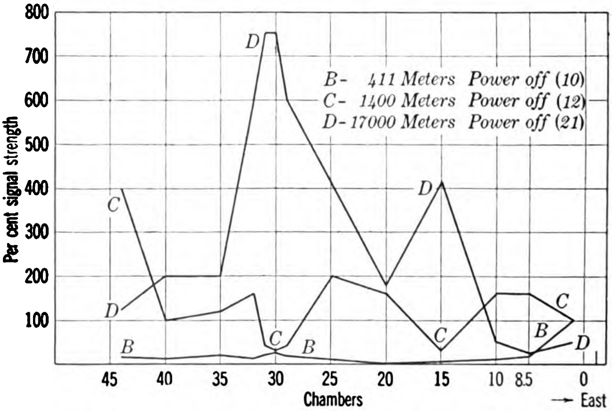

Fig. 21—Comparison of Signal Strengths, April 17, 1928.
Tunnel Blocked, Cables Grounded.

order to obtain the maximum signal strength. Whether this indicates that the transmission was coming along the wires or was merely being confined to a general path by the walls of the tunnel, is difficult to state with any certainty.

Conclusion

The results considered above are very far from being conclusive, but time did not permit of further tests being made. From a consideration of the curves already discussed, it would seem that the following general conclusions might be drawn,

(1) The penetration of any wave into the tunnel is some definite function of the frequency of the wave.

(2) Waves below 100 meters do not penetrate rock or soil to any appreciable extent.

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(3) Cables and rails do not conduct the shorter waves and broadcast waves to the same extent that they do long waves, although their effect on the broadcast band is appreciable.

(4) Wires and cables when ungrounded appear to act as wave antennas. When grounded they act as loops depending, of course, on the methods employed in carrying out the grounding.

(5) Much more energy appears to enter via the tunnel mouth than was at first suspected. This is indicated by the effect of the air shaft already discussed.

Dr. A. S. Eve, one of the co-workers in these experiments, has made the following comment,

> The most noteworthy point in the experiments is the fact that the short-wave signals (40 to 55 meters) did not penetrate well or far into the tunnel through any of the three possible channels:
>
> (1) through rock
> (2) through mouths and shaft
> (3) along conductors.
>
> The longer waves probably entered by all three methods, and Major Steel, who has prepared this report, is disposed from the evidence to attach considerable weight to the admission by air. If he is correct in this view there are interesting questions which arise connected with the nature of radio waves. Can they enter horizontal tunnels and also descend vertical shafts? Nor is it clear why the short waves should fail to go far into the mouth, while the long waves seem to penetrate the whole length of the tunnel.

Suggestions for Further Work

It would seem desirable to carry out further tests at some point where a tunnel exists without wires and rails leading into it. This may be difficult or, perhaps, impossible to carry out, but further experiments in the Mount Royal Tunnel would hardly seem advisable in view of the tremendous expense involved in completely screening or blocking the openings.

In future tests arrangements should be made to measure actual field strengths in microvolts per meter, some standard form of antenna being adopted for each particular wave. For the waves below 17000 meters it seems advisable to use a well constructed loop antenna.

In conclusion the writers wish to express their appreciation for the very great assistance given them by the President, Sir

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Henry Thornton, and by all the officials of the Canadian National Railways System with whom we came in contact. We particularly wish to mention the following: Mr. R. C. Johnson, Superintendent, Canadian National Railways; Mr. C. P. Price, Electrical Superintendent, Canadian National Railways; Mr. W. Walker, District Engineer, Canadian National Railways.

Dr. D. A. Keys and Dr. E. S. Bieler, of the Physics Department, McGill University, have been particularly helpful both in the preliminary tests and in the discussions on the results obtained. Great credit is also due to Mr. Rushbrook of the Canadian Marconi Company for his assistance in making the tests, and to the officials of the Department of National Defence for permission to use much of the apparatus required for the experiments.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

A NOTE ON THE DIRECTIONAL OBSERVATIONS ON GRINDERS IN JAPAN*

By

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(Electrotechnical Laboratory, Ministry of Communications, Tokyo, Japan)

BEFORE the General Assembly of the U. R. S. I., 1927, a paper entitled "The Directional Observations on Atmospherics in Japan" was presented by the authors, and the following conclusions had been arrived at by the series of observations.

(1) No appreciable change had been found in the observed directions of atmospherics by the variation of wavelengths which lie between 10,000 and 20,000 meters.

(2) On the diurnal observations from 7 a.m. to 9 p.m. no appreciable regular change had been noticeable in the observed directions of atmospherics by the time of day. Though some expected variations had, in general, been noticed at sunrise, sunset, and night times, they had not been so remarkable as in the case of signal observations.

(3) The directions of clicks had been obviously different from those of grinders, which seems to indicate that the seat of clicks differs from those of grinders.

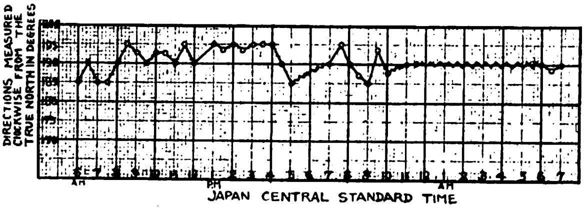

Fig. 1—February 17–18, 1928.
$\lambda = 10{,}000$ m.

(4) The diurnal variation had been much greater in the directions of clicks than in those of grinders. This seems to indicate that the seat of clicks lies nearer to the observed region than those of grinders, if the seats of both kinds of atmospherics might be assumed to have nearly equal areas. Judged from

* Dewey decimal classification: R 114. Original manuscript received by the Institute, September 7, 1928. Presented before the General Assembly of U.R.S.I., Brussels, Belgium, September, 1928.

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the great differences in the intensities, as well as in the nature of tones of clicks and grinders, this assumption may be considered to be probable.

(5) The directions of clicks had shifted from nearly South toward West as the season advanced from winter to summer, while those of grinders had remained unchanged.

(6) In summer, the clicks, which seem to originate from the mountain regions of Japan, seem to be most prevalent, while grinders probably come always from a tropical region such as the Dutch East-Indies.

Those were the results obtained from the observations over a period from February to July, 1927. The experiments were further continued and it was recognized that the same conclusions still hold good except in the night observations of grinders.

During the experiments, ten times of 24-hour observations were made covering all the seasons of the year. Though no other particular phenomena were noticed in the observations on clicks, it was remarkably found in those on grinders, on the contrary, that the directive property became indistinctive or disappeared at night. One example of directional variations (wavelength 10,000 meters) is shown in Fig. 1 in which the zig-zag lines indicate the durations of directivity fading, and in Fig. 2 which illustrates the change in the broadness of directivity in the same observation as shown in Fig. 1. In other 24-hour observations, the general tendencies of the variations were quite similar, though the times at both extremities of the fading bands of directivity shifted by a few hours according to the dates.

It is in general believed that the origins of long-wave atmospherics, as received in both England and the United States, are mainly in the lands situated in the Tropical Zone. [1, 2, 3] The grinders as received in Japan had also been indicated by the authors to be of a tropical origin such as the Dutch East-Indies. Taking the above into consideration and the daily movement of the illuminated and the darkened hemispheres into account the atmospherics originating at the tropical region in America seem very probable to be received in Japan as grinders early

[1] Lloyd Espenschied, C. N. Anderson, and Austin Bailey, "Transatlantic Radio Telephone Transmission," Proc. I.R.E., **14**, 7; February, 1926.

[2] R. A. Watson Watt, "The Directional Recording of Atmospherics," *Jour.* I.E.E., **64**, p. 596; May, 1926.

[3] L. W. Austin, "Direction Determinations of Atmospheric Disturbances on the Isthmus of Panama," Proc. I.R.E., **14**, 373; June, 1926.

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at night and those in Africa late at night, while those in the Dutch East-Indies may be received for all the time of day and night. Thus the grinders coming from other parts than the Dutch East-Indies seem to have made them apparently non-directive at night, because the unidirectional combination of antennas was

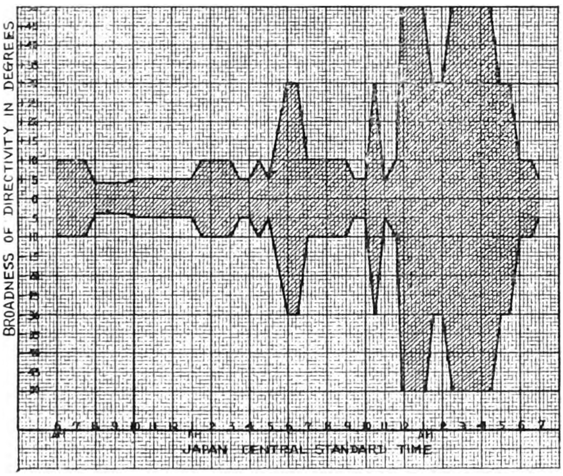

Fig. 2—February 17–18, 1928.
$\lambda = 10{,}000$ m.

not tried, a single loop being used as wave picking-up device in the observations. The observations are now still in progress, and it is expected that the phenomenon will be brought into light more definitely later on.

Acknowledgment is due to C. Asakawa and T. Takei of this aboratory for their assistance during the observations.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

ON THE BEHAVIOR OF NETWORKS WITH "NORMALIZED" MESHES*

By
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Summary—*The theory of normalizing meshes in electrical networks, as outlined in a recent article appearing in this journal,*[1] *is verified and illustrated by examples and figures relating to two- and three-mesh circuits. In the two-mesh circuit mesh No. 1 was normalized, thus confining the corresponding frequency to that mesh both for the transient and steady states as illustrated by Figs. 2 to 6 inclusive. For the three-mesh circuit meshes Nos. 1 and 3 were normalized, and the results are illustrated by Figs. 6 to 16 inclusive. In each case the theory was checked in every detail.*

In a recent issue of this journal[1] the mathematical theory concerning the interpretation of normal co-ordinates in dynamic systems as related to the electrical network was set forth in its more fundamental character, the stress being laid upon the analogy between a mechanical system of coupled mass points and its electrical network equivalent. At that time the interest lay chiefly in a deeper inquiry into this analogy which is so complete from the mathematical standpoint and yet so intangible from the physical angle.

The reason why such an investigation should present an unusual element of interest is due to the mechanical viewpoint rather than the electrical. For, in the mechanical system, the existence of so-called normal co-ordinates can easily be visualized, and these are known to possess distinct peculiarities; namely, of containing only one of the natural frequencies of the system. The projection of the oscillation of any mass point upon this particular direction in space will be simple harmonic, damped or undamped depending upon the presence or absence of frictional forces. This particular natural frequency is *isolated* in its corresponding normal co-ordinate in that it cannot be fed over into any other direction which is normal to it, or be brought into action unless the actuating force has a component in the direction of this normal co-ordinate.

* Dewey decimal classification: 621.319.2 Original manuscript received by the Institute, October 10, 1928.

[1] E. A. Guillemin. "Making Normal Co-ordinates Coincide with the Meshes of an Electrical Network," Proc. I. R. E., 15, 935; November, 1927.

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In pursuing the analogy between the mechanical and electrical systems it was soon recognized that the counterpart of a physical dimension or direction (co-ordinate) is a mesh in the equivalent electrical system. The question then immediately arose as to whether it would be possible in the electrical system to make a mesh correspond to what we recognize mathematically as being a normal co-ordinate, just as it is possible in the mechanical problem to orient our arbitrarily chosen reference axes so that

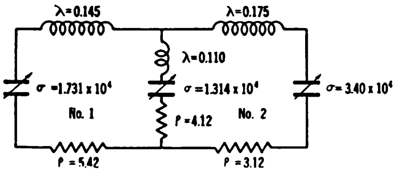

Fig. 1—Arrangement of Constants Used in Two-Mesh Circuit.

one of them will coincide with a normal direction. For, granting that this could be done, we should thus gain a means of isolating or confining one of the natural frequencies of the electrical network to that mesh. We should thus obtain a singular network which would show a distinctly different behavior, both in the

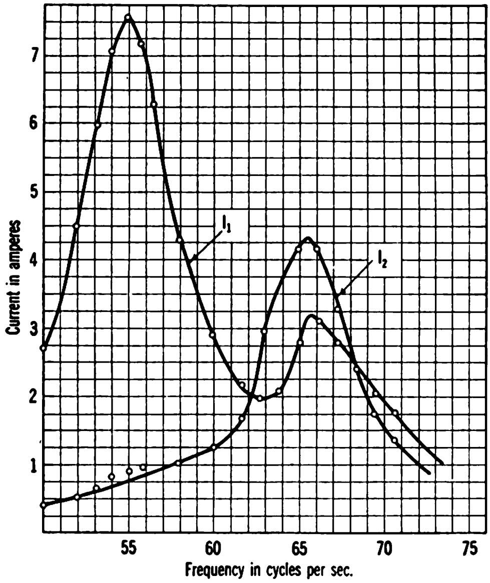

Fig. 2—Two-Mesh Circuit. Voltage in Mesh No. 1, Voltage Held Constant at 50.

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transient and steady states, from that of the ordinary network having the same number of natural frequencies. Certain transient frequencies would be absent in some of the meshes, or would not appear at all, depending upon the location of the disturbing force. Furthermore, since resonance phenomena depend upon coincidence between natural and impressed frequencies, the network would exhibit isolation properties with regard to resonance peaks in the steady state.

These and similar considerations made the investigation seem worth while from the practical standpoint, even though the problem was basically purely of mathematical interest.

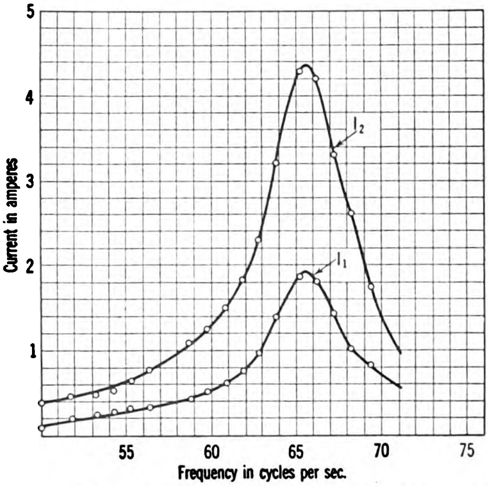

Fig. 3—Two-Mesh Circuit. Voltage in Mesh No. 2, Voltage Held Constant at 22.5.

It was for the purpose of providing concrete illustrations that an experimental investigation, of which the present paper is a report, was undertaken; and it is the hope of the writers that the following may prove interesting to those who have occasion to work with more intricate networks both theoretically and experimentally. For the fundamental theory as well as the notation used here we refer the reader to the paper already mentioned.

The first and simplest illustration will be given by the two-mesh circuit shown in Fig. 1. Each mesh contains independent inductance, resistance, and capacitance, and the two meshes are coupled by means of these three elements. Actually the common

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inductance shown in the center branch was a simple self-inductance of the air-core type, but it should be recognized that this might equally as well have been partly or wholly in the form of mutual inductance between two windings in proximity to each other, since mutual and self-inductance play exactly the same role both physically and mathematically.

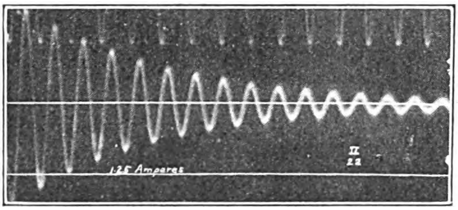

Fig. 4—Oscillogram for Two-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 1. Mesh 1 having been made a normal coordinate, the curve should be a damped double frequency. The beat shows this clearly. 60-cycle timing wave.

In this circuit the following constants were used:

$$\lambda_{11} = 0.255 \text{ henries}$$
$$\sigma_{11} = 3.045 \times 10^4 \text{ darafs}$$
$$\rho_{11} = 9.54 \text{ ohms}$$

$$-\lambda_{12} = 0.110 \text{ henries}$$
$$-\sigma_{12} = 1.314 \times 10^4 \text{ darafs}$$
$$-\rho_{12} = 4.12 \text{ ohms}$$

$$\lambda_{22} = 0.285 \text{ henries}$$
$$\sigma_{22} = 4.714 \times 10^4 \text{ darafs}$$
$$\rho_{22} = 7.24 \text{ ohms}$$

The system of differential equations for the case of no impressed forces becomes:

$$\lambda_{11}\frac{d^2x_1}{dt^2} + \rho_{11}\frac{dx_1}{dt} + \sigma_{11}x_1 + \lambda_{12}\frac{d^2x_2}{dt^2} + \rho_{12}\frac{dx_2}{dt} + \sigma_{12}x_2 = 0$$

$$\lambda_{12}\frac{d^2x_1}{dt^2} + \rho_{12}\frac{dx_1}{dt} + \sigma_{12}x_1 + \lambda_{22}\frac{d^2x_2}{dt^2} + \rho_{22}\frac{dx_2}{dt} + \sigma_{22}x_2 = 0$$

where:

$$x_1 = \int i_1 dt \quad \text{and} \quad x_2 = \int i_2 dt.$$

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Assuming the solutions:

$$x_1 = Y_1 e^{pt} \quad \text{and} \quad x_2 = Y_2 e^{pt},$$

we get:

$$b_{11}Y_1 + b_{12}Y_2 = 0$$

$$b_{21}Y_1 + b_{22}Y_2 = 0$$

with:

$$b_{11} = 0.255p^2 + 9.54p + 3.045 \times 10^4$$

$$b_{12} = b_{21} = -0.110p^2 - 4.12p - 1.314 \times 10^4$$

$$b_{22} = 0.285p^2 + 7.24p + 4.714 \times 10^4.$$

From inspection we see that:

$$\frac{b_{12}}{b_{11}} = \alpha = -0.432.$$

The value of α is insignificant. The fact that such a relation exists is the necessary and sufficient condition for making mesh No. 1

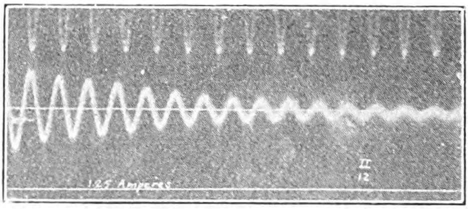

Fig. 5—Oscillogram for Two-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 2. Note that this is a damped single frequency. 60-cycle timing wave.

a normal co-ordinate. This is clear from the fact that the determinantal equation for this case becomes:

$$D = \begin{vmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{vmatrix} = 0$$

$$= \begin{vmatrix} b_{11} & \alpha b_{11} \\ \alpha b_{11} & b_{22} \end{vmatrix} = b_{11}b_{22} - \alpha^2 b_{11} = 0$$

or:

$$D = b_{11}(b_{22} - \alpha^2 b_{11}) = 0,$$

from which:

$$b_{11} = 0; \quad \text{or} \quad (b_{22} - \alpha^2 b_{11}) = 0.$$

The first of these determines that frequency which is confined to mesh No. 1. We shall call it the *normalized frequency*. The second

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of the above equations gives us that frequency which is not confined but still common to both meshes. We shall call it the *system frequency.*

Substituting numerical values, we get for the equation which determines the normalized frequency:

$$0.255p^2+9.54p+3.045\times 10^4=0$$

Since the circuit is highly oscillatory, the frequency is given very nearly by neglecting the resistance term (linear term in p). We have:

$$0.255p^2+3.045\times 10^4=0$$

or:

$$p=\pm j345=\pm j2\pi f,$$

so that the normalized frequency becomes:

$$f=55 \text{ cycles per second.}$$

In exactly the same way the system frequency is found to be 66.3 cycles per second. It should be borne in mind that although the above procedure of neglecting the resistance term is sufficiently accurate for the determination of the frequencies and a consequent discussion of steady-state phenomena, the damping constant which depends directly upon the resistance term becomes of vital importance in the consideration of transient phenomena. Since these facts are, however, well known in circuit theory, we shall not dwell upon them here.

In the actual experiment, the desired frequencies were chosen first and the constants of the circuit as well as the value of α adjusted so as to meet the available values of inductance and capacitance in the laboratory. Hence the odd figures given above. This method of reversing the calculations, which is essential in every problem of design, is given in a thesis[2] report filed in the library of the Massachusetts Institute of Technology and will not be reproduced here. We prefer rather to show the experimental results and give a discussion of them in the light of the theory upon which they are based.

Let us review the situation once more and see what results we should expect to obtain. We have made mesh No. 1 a normal co-ordinate to the frequency of 55 cycles per second. The remaining system frequency is 66.3 cycles per second. This

[2] "Experimental Verification of Theories Concerning Normalized Meshes in Electrical Networks," M. I. T. thesis, 1928, by W. Glendinning.

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means that the frequency 55 is confined to mesh No. 1, but that the frequency 66.3 is still common to both meshes. If we impress an alternator in mesh No. 1 and vary the frequency over a range covering the above values, then the current in mesh No. 1 will show a peak for 55 cycles and another for 66.3 cycles. The current in mesh No. 2 will show a peak only for 66.3 cycles. It cannot show a peak for 55 cycles because this frequency is no longer a natural frequency to that mesh but is confined entirely to mesh

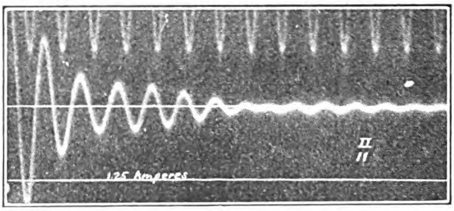

Fig. 6—Oscillogram for Two-Mesh Circuit. Voltage in Mesh 2, Current in Mesh 2. Note that this is a damped single frequency. 60-cycle timing wave.

No. 1. On the other hand, if we impress an alternator in mesh No. 2 and vary the frequency over the same range, we shall find that the currents in both meshes show peaks only at the system frequency of 66.3 cycles. The resonance peak corresponding to the normalized frequency of 55 cycles is suppressed even in the normal mesh (mesh No. 1)! This rather striking phenomenon is due to the fact already mentioned that the normalized frequency is so completely isolated within the corresponding normal mesh that it cannot be brought into action unless the actuating force is impressed in that mesh.

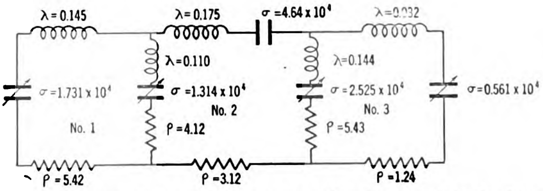

Fig. 7—Arrangement of Constants for Three-Mesh Circuit.

As to the transients that we have to expect we can predict that if we suddenly impress a constant emf in either mesh and take an oscillogram of the resulting current in the *other* mesh, then we should find that the result is a *simple damped harmonic* of system frequency. The same is true of the current in the

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non-normal mesh (mesh No. 2) with the emf impressed in No.2. However, if we examine the current in mesh No. 1 with the emf impressed there, then we shall find a superposition of two damped frequencies—the normalized and the system frequencies.

Figs. 2 and 3 show the experimental results of the steady-state conditions, and bear out the theory in every detail.

For the transient results given by the oscillogram Figs. 4, 5, and 6, the roman II designates the two-mesh circuit, while the 11, 12, and 22 indicate that current is measured in mesh No. 1 with voltage in No. 1, or current measured in No. 1 with voltage

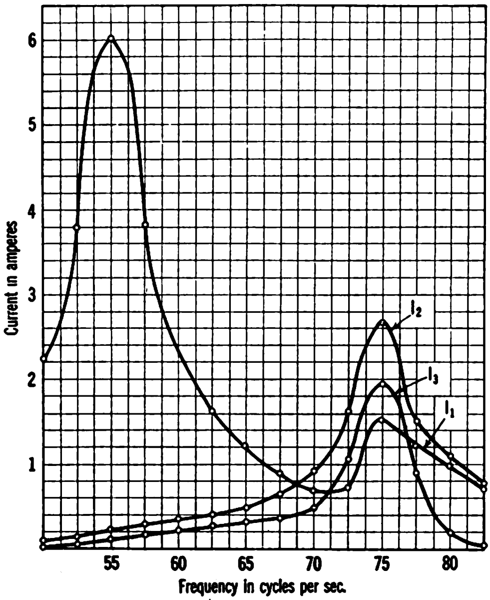

Fig. 8—Three-Mesh Circuit. Voltage in Mesh No. 1, Voltage Held Constant at 40.

in No. 2 or vice versa, or current measured in No. 2 with voltage in No. 2. With this in mind the figures need no further explanatory note. The same system of notation will be extended to the three-mesh circuit to be discussed later.

Close inspection of Figs. 4, 5, and 6 will show that the one designated 11 is the only one which contains two frequencies superimposed as may be detected from the beat. The other two are simple damped sinusoids.

A second study was made, using the three-mesh circuit shown in Fig. 7 with the constants so chosen that meshes Nos. 1 and 3

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become normal co-ordinates. The values of the constants were fixed as follows:

$$\lambda_{11} = 0.255 \text{ henries}$$
$$\rho_{11} = 9.54 \text{ ohms}$$
$$\sigma_{11} = 3.045 \times 10^4 \text{ darafs}$$

$$-\lambda_{12} = 0.110 \text{ henries}$$
$$-\rho_{12} = 4.12 \text{ ohms}$$
$$-\sigma_{12} = 1.314 \times 10^4 \text{ darafs}$$

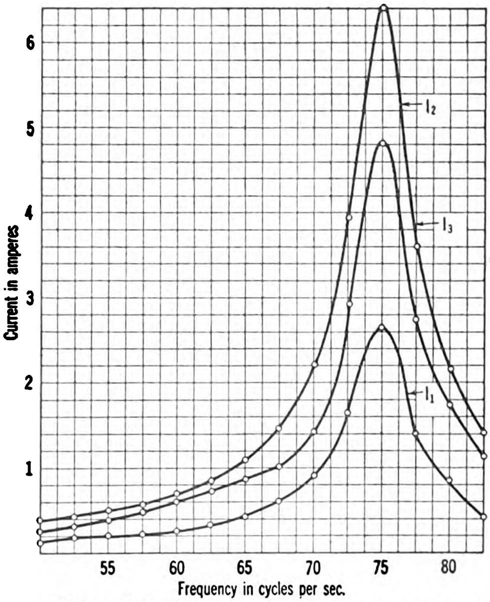

Fig. 9—Three-Mesh Circuit. Voltage in Mesh No. 2, Voltage Held Constant at 40.

$$\lambda_{22} = 0.429 \text{ henries}$$
$$\rho_{22} = 12.67 \text{ ohms}$$
$$\sigma_{22} = 8.479 \times 10^4 \text{ darafs}$$

$$-\lambda_{23} = 0.144 \text{ henries}$$
$$-\rho_{23} = 5.43 \text{ ohms}$$
$$-\sigma_{23} = 2.525 \times 10^4 \text{ darafs}$$

$$\lambda_{33} = 0.176 \text{ henries}$$
$$\rho_{33} = 6.67 \text{ ohms}$$
$$\sigma_{33} = 3.086 \times 10^4 \text{ darafs}$$

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Thus we obtained for the three-mesh circuit:

$$b_{11}=0.255p^2+9.54p+3.045\times 10^4$$
$$b_{12}=-0.110p^2-4.12p-1.314\times 10^4$$
$$b_{22}=0.429p^2+12.67p+8.479\times 10^4$$
$$b_{23}=-0.144p^2-5.43p-2.525\times 10^4$$
$$b_{33}=0.176p^2+6.67p+3.086\times 10^4.$$

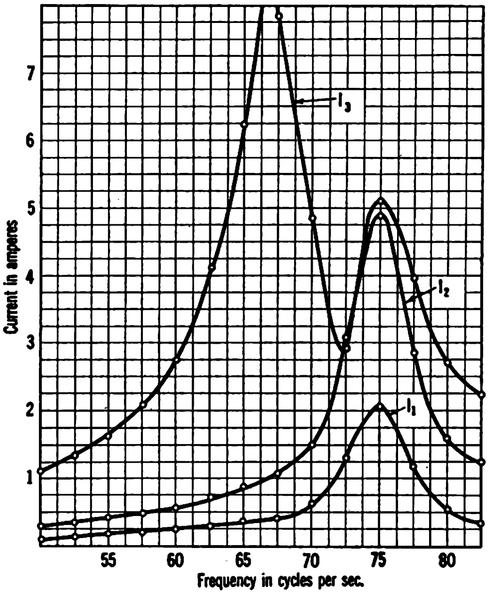

Fig. 10—Three-Mesh Circuit. Voltage in Mesh No. 3, Voltage Held Constant at 40.

Here we see that:

$$\frac{b_{12}}{b_{11}}=\alpha=-0.432; \quad \text{and} \quad \frac{b_{23}}{b_{33}}=\beta=-0.818.$$

Again the numerical values of α and β are not significant. The fact that the same ratio exists between inductance, resistance, and capacitance, is the necessary condition for making a mesh a normal co-ordinate. Here we have made meshes Nos. 1 and 3 normal co-ordinates.

The determinantal equation for this case becomes:

$$D(p)=\begin{vmatrix} b_{11} & \alpha b_{11} & 0 \\ \alpha b_{11} & b_{22} & \beta b_{33} \\ 0 & \beta b_{33} & b_{33} \end{vmatrix}=0$$

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or:

$$b_{11}b_{33}(b_{22}-\beta^2 b_{33}-\alpha^2 b_{11})=0,$$

which is factorable into:

$$b_{11}=0; \quad b_{33}=0; \quad (b_{22}-\beta^2 b_{33}-\alpha^2 b_{11})=0.$$

The fact that b_{11} and b_{33} can be factored out in this way shows that meshes Nos. 1 and 3 are normal co-ordinates. The corresponding normal frequencies are given by the roots of the first two of the above equations. Incidentally this process of making

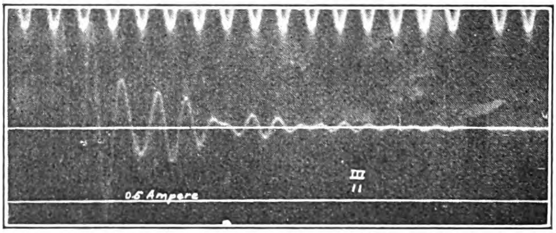

Fig. 11—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 1. Note that this is a damped double frequency as can be seen from the beat. 60-cycle timing wave.

meshes correspond to normal co-ordinates makes it easy to solve for the roots or natural frequencies. Ordinarily the three-mesh circuit discussed here would lead to a sixth-degree equation with three pairs of conjugate complex roots. This is true in our case also, but the equation is factorable into three quadratics which

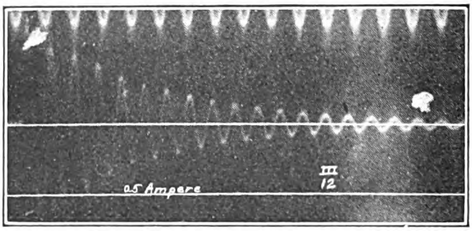

Fig 12—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 2. Note that this is a damped single frequency. The normal frequency of mesh 1 is suppressed and does not appear in mesh 2. 60-cycle timing wave.

may be solved by the simple formula. The investigation of networks with normalized meshes thus becomes much simpler than it is for ordinary networks. Also the evaluation of integration constants or transient current amplitudes by Heaviside's formula becomes very simple; but we cannot go into that here.

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We find by solving the three quadratics given above that the normalized frequency belonging to mesh No. 1 is 55 cycles per second, that belonging to mesh No. 3 is 66.7 cycles per second, and that the remaining frequency is 75 cycles.

Here, as in the case of the two-mesh circuit, the values of frequency were chosen to start with and the constants as well as values of α and β adjusted to meet the available values. For this process of adjustment or "design calculation" we again refer to the thesis already mentioned above.

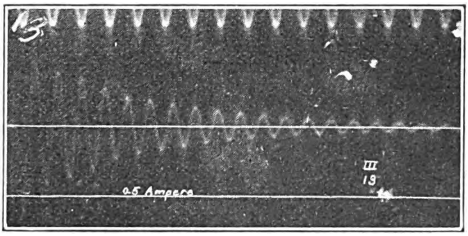

Fig. 13—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 1, Current in Mesh 3. Note that this is a damped single frequency. The normal frequency of mesh 1 is confined to that mesh. 60-cycle timing wave.

Let us now see what we should expect in the way of transient and steady-state results for this three-mesh network with meshes Nos. 1 and 3 normalized. Suppose we impress an alternator in mesh No. 1 and—keeping the voltage constant—vary the frequency so as to cover from 55 to 75 cycles per second. We should

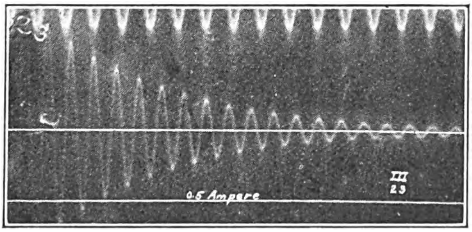

Fig. 14—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 2, Current in Mesh 2. Note that this is a damped single frequency. Mesh 2 is a non-normal mesh and only the system frequency will appear. The normal frequencies of meshes 1 and 2 do not come into action. 60-cycle timing wave.

expect the current in mesh No. 1 to show a peak at 55 cycles—its normal frequency—and another peak at 75 cycles—the system frequency. No peak should occur at 66.7 cycles because this is the

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frequency which is isolated in mesh No. 3. For the current in mesh No. 2 we should expect only one peak at 75 cycles, and for the current in mesh No. 3 we should expect only *one* peak also at 75 cycles. We do *not* get a peak for 66.7 cycles—the normalized frequency for mesh No. 3—because the driving force is not in that mesh but is located in mesh No. 1. This is a very interesting fact; namely, that a response to a normalized frequency can be brought about only by applying the emf in that mesh to which the normalized frequency in question is confined. With the generator in mesh No. 1, as we have here assumed, meshes Nos. 2 and 3 have only one resonance frequency. These facts were borne out very nicely in our experiments, and are shown in Fig. 8.

Suppose now we consider the generator in mesh No. 2 and again vary the frequency as before, keeping the voltage constant. In this case the currents in all three meshes should show only *one* resonance peak at 75 cycles—the system frequency. The

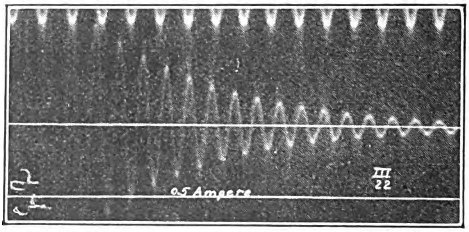

Fig. 15—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 2, Current in Mesh 3. Note that this is a damped single frequency. 60-cycle timing wave.

normalized frequencies of 55 and 66.7 cycles in meshes Nos. 1 and 3 do not respond, and this is again clearly shown by our experimental curves of Fig. 9 for this case.

Finally, suppose we impress the voltage in mesh No. 3 and make the same run as above. We expect in meshes Nos. 1 and 2 a response only to the system frequency of 75 cycles because the normalized frequency of 55 cycles in mesh No. 1 does not come into action. In mesh No. 3, however, we now expect to find a response to 66.7 cycles—its normalized frequency—and also to the system frequency of 75 cycles. These conditions were also verified experimentally and are illustrated by Fig. 10.

Now as to the type of transients we should expect to get. Here, as in the two-mesh case, we use the numbering 12, 13, etc., to designate runs, whereby the first figure designates the mesh

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in which the current is observed and the second, the mesh in which the constant emf is suddenly applied, or vice versa. Thus the runs 12, 13, 22, and 23 should show simple damped sinusoids of system frequency. The run 11, on the other hand, should show a double-frequency transient containing 55 and 75 cycles, and the run 33 should also show a double-frequency transient containing 66.7 and 75 cycles. The oscillogram for run 11 is given in Fig. 11. Those for runs 12, 13, 22, and 23 are given in Figs. 12, 13, 14, and 15, respectively, while the oscillogram for run 33 is given in Fig. 16. The double frequency of runs 11 and 33 is detected in each case by the beat, which is absent in the oscillograms for the remaining runs.

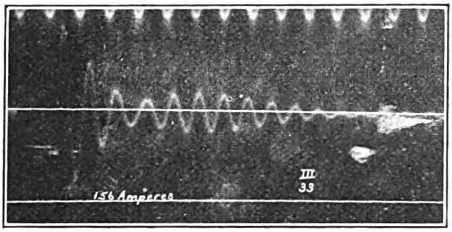

Fig. 16—Oscillogram for Three-Mesh Circuit. Voltage in Mesh 3, Current in Mesh 3. Mesh 3 having been made a normal coordinate, the curve is a damped double frequency. The beat shows this clearly. 60-cycle timing wave.

We see that our mathematical theory is substantiated very nicely by these simple experiments, and it is safe to assume that the theory holds in its general form and applies to any type of network no matter how complicated. It seems reasonable that these properties of isolation and suppression of resonance effects should have some practical application in the design of circuits with special characteristics, and it is the hope of the authors that the principles involved have been made sufficiently clear by these illustrations so that more general interest will be aroused. In the latter event we earnestly invite criticism and co-operation.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 2 *February, 1929*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE*

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. Radio Communication

R007.1 Dellinger, J. H. Analysis of broadcasting station allocation. Proc. I. R. E., **16**, 1477-85; November, 1928.

(A discussion of the technical aspects of the reallocation of November 11, 1928.)

R100. Radio Principles

R113 Diagramme des champs élèctriques mésures a Meudon pendant le deuxieme semestre 1927. (Diagrams of electric fields measured at Meudon during the second half of 1927). *L'Onde Electrique*, **7**, 458-460; October, 1928.

(Curves taken on LY, UA, WSS, and GBL during second half of 1927.)

R113.1 Colwell, R. C. Fading curves along a meridian. Proc. I. R. E., **16**, 1570-1573; November, 1928.

(Fluctuations in signal strength of KDKA, Pittsburgh, Pa. were observed through the sunset period of Morgantown, W. Va. Observations made covered twenty-one days. On bright clear days, the curve fluctuated considerably while on cloudy days the curve was fairly steady.)

R113.1 Merritt, E. and Bostwick, W. E. A visual method of observing the influence of atmospheric conditions on radio reception. *Proc.* Nat. Acad. Sci., **14**, 884-88; November, 1928.

(A method is described which utilizes the cross-coil system and a cathode-ray oscillograph. It was possible to notice visually several successive rotations of the plane of polarization of the down-coming wave during sunset period.)

R113.4 Størmer, C. Short-wave echoes and the Aurora Borealis (letter). *Nature* (London), **122**, 681; Nov. 3, 1928.

(Report on signals and echoes received at Bygdø, Oslo from the short-wave station at Eindhoven, Holland. The echoes arrived from 3 to 15 seconds later than the principal signal. The writer explains these belated echoes by the theory that radio waves penetrate the Heaviside layer passing then into empty pockets of large dimensions. The pockets are surrounded by walls of electrons from which the waves are reflected.)

* Original manuscript received by the Institute, December 14, 1928.

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R113.4 Radio echoes and magnetic storms (letters). *Nature* (London), 122, 768; November 17, 1928.

(Two notes, one by S. Chapman and the other by E. T. Eckersley with respect to the note (*Nature*, p. 681, Nov. 3, 1928) by Størmer on the explanation for the long time interval echoes. Eckersley calls attention to his paper in the *Philosophical Magazine*, June, 1925, where he explained the whistles of lowering pitch by means of dispersion and the group velocity along the path in the Heaviside layer. Chapman calls attention to the fact that the positive ions are also present in addition to the electrons which were considered in Størmer's theory.)

R113.4 Beatty, R. T. Short wave signals which travel over a million miles and return. *Wireless World and Radio Review*, **23**, 722-23; Nov. 28, 1928.

(A picturization of electron vorteces of extremely large dimensions and their possible effects on radio transmission explaining long time interval echoes.)

R113.4 Schelleng, J. C. Note on the determination of the ionization in the upper atmosphere. Proc. I. R. E., **16**, 1471-1476; November, 1928.

(Describes a method of estimating distribution of ionization in upper atmosphere This method based on measurements on several frequencies of effective height as determined by interference or echo experiments.)

R116 Bergmann, L. and Holzlöhner, G. Über die Fortpflanzungschwindigkeit elektrischer Wellen an dunnen Drähten von verschiedenem Leitvermögen. (On the velocity of propagation of electric waves along thin wires of different conductivity.) *Ann. d. Phys.*, **87**, 653-676; No. 21, 1928.

(Brief review of the theories of propagation by Hertz, Rayleigh, Drude, Mie, Sommerfeld, Mercier, Hund, and experimental data with thin wires of copper, aluminium, brass, German silver, manganin, and constantan.)

R145.3 Bashenoff, V.I. Supplementary note to abbreviated method for calculating the inductance of irregular plane polygons of round wire. Proc. I. R. E., **16**, 1553-1558; November, 1928.

(Additional discussion based on this paper.)

R200. Radio Measurements and Standardization

R210 Rangachari, T. S. A method of calibrating a low frequency valve generator with a single frequency standard source. *Experimental Wireless and Wireless Engineer* (London), **5**, 633-634; November, 1928.

(The source of unknown frequency is connected across a series combination of resistance and an inductance and the former is varied until the voltage across it is the same as across the inductance. The same is done when an emf of known frequency is impressed. The unknown frequency can then be calculated from resistance settings and the known frequency.)

R214 Hitchcock, R. C. Quartz crystals: How to cut and grind them. *Radio Broadcast*, **14**, 85-87; December, 1928.

(Formulas for the thickness vibration are given for the 30-degree and the Curié cut. The method of cutting and grinding piezo-electric quartz plates is described.)

R214 Harrison, J. R. Piezo-electric oscillator circuits with four-electrode tubes. Proc. I.R.E., **16**, 1455-1470; November, 1928.

(Description of two piezo-oscillator circuits which use screen-grid vacuum tubes. One circuit utilizes two pairs of electrodes and feedback takes place through the crystal. In the other circuit, the two electrodes of the crystal are connected between the plate and the control grid.)

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R214 Terry, E. M. The dependence of the frequency of quartz piezo electric oscillators upon circuit constants. Proc. I. R. E., **16**, 1486-1506; November, 1928.

(Mathematical theory of the piezo-electric quartz oscillator is given. It treats the case where the quartz element is between the grid and the filament as well as the case when between the grid and the plate. The analysis is made for the tuned-plate circuit, inductance and resistance load.)

R214 Strout, R. S. Temperature coefficient of quartz crystal oscillators. *Phys. Rev.*, **32**, 829-831; November, 1928.

(The temperature coefficient of the frequency of a quartz plate 1.8×1.8×0.11 oscillating with the thickness frequency was found to decrease linearly from 22.7 parts per million per degree at 65 deg. C to 1.6 parts per million per degree at—189 deg. C.)

R280 Rybner, J. Note sur les experiences relatives aux proprietes dielectriques des gaz ionises de MM. Gutton et Clement. (Note on the experimental results of dielectric properties of ionized gas found by Gutton and Clement). *L'Onde Electrique*, **7**, 428-436; October, 1928.

(Shows that the experimental facts observed originally by Gutton and Clement can be explained by the classical theory and that it is not necessary to assume that the ions are not free and their movement due to quasi elastic forces.)

R300. Radio Apparatus and Equipment

R330 Millen, J. The control grid glow tube. *Radio Engineering*, **8**, 18-19; September, 1928.

(Description of this tube giving the breakdown characteristic and other useful data.)

R331 Van Dyck, A. F. and Engel, F. H. Vacuum tube production tests. Proc. I. R. E., **16**, 1532-52; November, 1928.

(General description is given of methods and apparatus for testing tubes in large quantities.)

R343 Van Dyck, A. F. and Dickey, E. T. Quantitative methods used in tests of broadcast receiving sets. Proc. I. R. E., **16**, 1507-1531; November, 1928.

(General classes of receiving set measurement used by authors are outlined as special engineering tests and production tests. New form of radio-frequency generator designed for this type of work is described.)

R343.7 Warren, A. G. Surges in eliminator smoothing circuits. *Experimental Wireless and Wireless Engineer* (London), **5**, 601-606; November, 1928.

(A discussion of the effects of surges occurring in an eliminator circuit which is supplied from d. c. mains. It is shown that the effects are most serious when the set is not connected or the filaments of the set are not burning.)

R344 Aisberg, E. Utilisation des lampes de T.S.F. pour la production de musique electrique. (Use of vacuum tubes for the production of electric music). *L'Onde Electrique*, **7**, 455-458; October, 1928.

(Reviews the various systems for producing music by means of vacuum-tube generators.)

R344 A tuning fork controlled audio oscillator. *Jnl. Scien. Instr.* (London), **5**, 361-362; November, 1928.

(Description of audio oscillator made by General Radio Company.)

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R385.5 Meyer, E. Über eine einfache Methode der automatischen Klanganalysis und der Messung der Nichtlinearitat von Kohlemicrophonen. (A simple method of automatic sound analysis and measurement of distribution in carbon microphones). *Elektrische Nachrichten Technik*, **5**, 398-403; October, 1928.

(A bridge method is described for testing carbon microphones. The method is a simplified arrangement for sound analysis.)

R388 Coursey, P. R. A portable oscillograph equipment for observation of transient electrical phenomena. *Experimental Wireless and Wireless Engineer* (London), **5**, 616-619; November, 1928.

(Description of the apparatus.)

R400. Radio Communication Systems

R470 Fuller, L. F. and Tolson, W. A. Power line carrier telephony. *Jnl.* A. I. E. E., **47**, 711-715; October, 1928.

(Outlines briefly modern communication requirements, provision of communication and different types of equipment available for this service.)

R470 Wolfe, W. B. and Sarros, J. D. Problems in power line carrier telephony and recent developments to meet them. *Jnl.* A. I. E. E., **47**, 727-731; October, 1928.

(Outline of difficulties met in application of high-frequency communication to power systems. Description of recent developments in equipment.)

R500. Applications of Radio

R520 Craft, E. B. Airways communication service. *Bell System Tech. Jour.*, **7**, 797-807; October, 1928.

(Two-way radio telephone communication between airplane and ground experiments by the A. T. and T. Co. Describes in addition the system used by the U. S. Government and between London and Paris. Suggests that further channels in the high-frequency band should be assigned to this work.)

R531.2 Radio stations of the world on frequencies above 1500 kc. Proc. I. R. E., **16**, 1575-1604; November, 1928.

(List of high-frequency stations throughout the World.)

R551 Bigourdan, G. Sur l'unification des signaux horaires radio telegraphiques. (On the unification of radio time signals). *Rev. Generale d'electricité*, **24**, 702-703; November 10, 1928.

(Chart for time signals.)

R582 Scheppmann, W. and Eulenhöfer, A. Der Bildfunk nach dem System Lorenz-Korn. (Picture transmission system of Lorenz-Korn). *Elektrische Nachrichten Technik*, **5**, 373-381; October, 1928.

(Describes sending and receiving equipment as well as the latest development in synchronization.)

R582 Oglobinski, M. G. Derniers progress de la transmission Belinographique en France. (Latest progress on the transmission of pictures by the Belin system). *L'Onde Electrique*, **7**, 446-455; October, 1928.

(Description of the improved Belin system using photoelectric tubes.)

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R584 Hermann, P. W. High frequency electric furnaces. *Radio* (San Francisco), **10**, 12-13; December, 1928.

(Methods and equipment used for melting precious metals.)

R800. Non-Radio Subjects

535.3 Bergstein, M. Applications of the photoelectric cell in industry. *Radio Engineering*, **8**, 28-30; September, 1928.

(Uses in commercial work.)

538 Maurain, Ch. Activite solaire et magnetisme terrestre. (Solar activity and terrestrial magnetism). *L'Onde Electrique*, **7**, 414-427; October, 1928.

(Review of the solar activity and its relation to the diurnal variations. Includes a list of references.)

621.314.6 Turner, H. M. The constant impedance method for measuring inductance of choke coils. Proc. I. R. E., **16**, 1559-1569; November, 1928.

(Description of a method of measuring at power frequencies the inductance of iron-cored choke coils of type commonly used in radio as filters.)

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VOLUME 17 MARCH, 1929 NUMBER 3

PROCEEDINGS
of
The Institute of Radio Engineers

1929 CONVENTION
Washington, D. C.
May 13-15

General Information and Subscription Rates on Page 404

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Institute of Radio Engineers

Forthcoming Meetings

FOURTH ANNUAL CONVENTION
Washington, D. C., May 13–15, 1929

ATLANTA SECTION
Atlanta, Georgia, March 6, 1929

DETROIT SECTION
Detroit, Mich., March 15, 1929

LOS ANGELES SECTION
Los Angeles, Calif., March 18, 1929

NEW YORK MEETING
New York, N. Y., April 3, 1929

PHILADELPHIA SECTION
Philadelphia, Penna., March 22, 1929

PITTSBURGH SECTION
Pittsburgh, Penna., March 19, 1929

SAN FRANCISCO SECTION
San Francisco, Calif., March 20, 1929

WASHINGTON SECTION
Washington, D. C., March 14, 1929

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 17 March, 1929 Number 3

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The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

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Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership etc., are priced at seventy-five cents per copy per year.

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Entered as second class matter at the Post Office at Menasha, Wisconsin.

Acceptance for mailing at special rate of postage provided for in the Act of February 28, 1925, embodied in paragraph 4, Section 412, P. L. and R. Authorized October 26, 1927.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered, and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer de visions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportions.

Abbreviations—Write a.c. and d.c., kc, μf, μμf, emf, mh, μh, henries, abscissas, antennae Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right, in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

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Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

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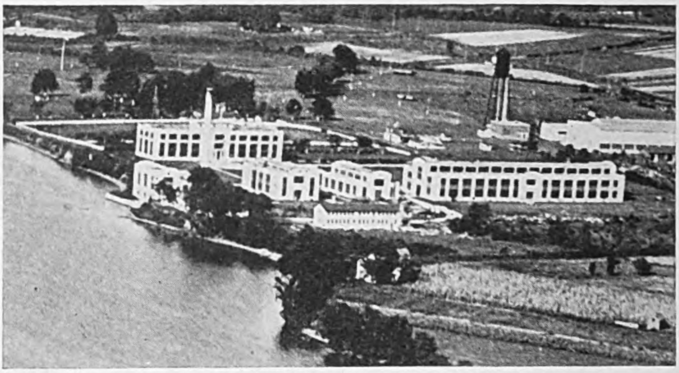
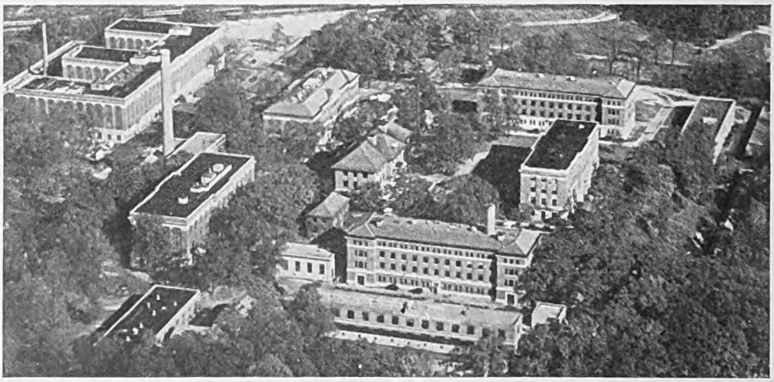


General View of Naval Research Laboratory, Anacostia, D. C.
(Several additional buildings have been added since this photo was taken.)



Airplane View of Bureau of Standards, Washington, D. C.
Radio Building at right in background.

INSPECTION TRIPS AT CONVENTION

Two full afternoons of the Fourth Annual Convention will be devoted to inspection trips to the Naval Research Laboratory and the Bureau of Standards, in Washington.

The Naval Research Laboratory is located on the Potomac River about eight miles from the center of Washington. It will be reached by buses going via the White House, Treasury Building, up historic Pennsylvania Avenue, past the Capitol, Botanical Gardens, Congressional Library, Navy Yard, and St. Elizabeth's Hospital.

The trip to the Bureau of Standards will be made by buses also. The route is to include the fine residential section of Washington, and a four-mile ride through Rock Creek Park. Returning, the convention delegates will cross the Million Dollar Bridge, past the former home of Herbert Hoover, the "S" Street home of Woodrow Wilson, and many other historic places of interest.

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INSTITUTE NEWS AND RADIO NOTES

February Meeting of the Board of Direction

The February meeting of the Board of Direction of the Institute, held on the sixth of the month in the offices of the Institute in New York City, was attended by the following Board members: A. Hoyt Taylor, President; Melville Eastham, Treasurer; Arthur Batcheller, R. A. Heising, J. V. L. Hogan, L. M. Hull, C. M. Jansky, Jr., R. H. Manson, L. E. Whittemore, and J. M. Clayton, Secretary.

The following were transferred or elected to higher grades in the Institute:

Transferred to the Member grade, Harold Gray; elected to the Member grade, Lieut. H. N. Coulter and Mrs. J. D. Stewart.

One hundred and six Associate members and ten Junior members were elected.

With great regret the Board of Direction acceded to Dr. Goldsmith's request that he should not be reelected Editor of Institute Publications, a position which he has held for the past sixteen years. By unanimous vote of those members present the following resolution was adopted by the Board:

> WHEREAS: The members of the Board of Direction of the Institute of Radio Engineers are deeply appreciative of the invaluable services rendered by
>
> **Alfred Norton Goldsmith**
>
> as Editor of the PROCEEDINGS since the formation of the Institute; and
>
> WHEREAS: This contribution made by him as a pioneer in the development of the radio art is immeasurable in the light of future progress; and
>
> WHEREAS: The high standards and ethics adopted by him as Editor have earned for the PROCEEDINGS an enviable position in the field of radio engineering literature;
>
> *Therefore Be it Resolved:* That the Board of Direction, as individuals and as a whole, express to Dr. Goldsmith deep regret at the termination of his tenure of office as Editor, and profound appreciation of his untiring efforts and conscientious zeal, to which the PROCEEDINGS for the past sixteen years shall always remain a fitting testimony.

To succeed Dr. Goldsmith as Editor of Publications and Chairman of the Board of Editors, the Board appointed Dr. Walter G. Cady of Scott Laboratory, Wesleyan University,

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Middletown, Connecticut. Dr. Cady served on the Board of Direction for 1928 and was the recipient of the 1928 Morris Liebmann Memorial Prize. He is well known as a professor of physical science, a research worker in piezo electricity, a past associate editor of the *Physical Review*, and a contributor on numerous occasions to the PROCEEDINGS.

Radio Stations of Holland

From the Dutch Government a revised list of high-frequency assignments in Holland (1,500 kc and above) has been received by the Institute office. This is a revision of the information relating to Dutch stations contained in the list of high-frequency stations published in the November, 1928, issue of the PROCEEDINGS. Copies of the revision are available to members upon request at the Institute office.

Standard Frequency Transmissions by the Bureau of Standards

The Bureau of Standards announces a new schedule of radio signals of standard frequencies, for use by the public in calibrating frequency standards and transmitting and receiving apparatus. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington which went into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to 1,000 miles from the transmitting station.

The transmissions are by continuous-wave radiotelegraphy. The modulation which was previously on these signals has been eliminated. A complete frequency transmission includes a "general call" and "standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues for about 2 minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about 4 minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute in-

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terval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in the Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

Radio Transmissions of Standard Frequency; Schedule of Frequencies in Kilocycles

Eastern Standard Time	March 20	April 22	May 20	June 20	July 22
10:00–10:08 P. M.	1500	4000	125	550	1500
10:12–10:20	1700	4500	150	600	1700
10:24–10:32	2250	5000	200	700	2000
10:36–10:44	2750	5500	250	800	2300
10:48–10:56	2850	6000	300	1000	2700
11:00–11:08	3200	6500	375	1200	3100
11:12–11:20	3500	7000	450	1400	3500
11:24–11:32	4000	7300	550	1500	4000

Changes in Federal Radio Commission

The Institute is honored in the action of the President of the United States in appointing to the Federal Radio Commission two of its Board members:

Arthur Batcheller and C. M. Jansky, Jr.

At this writing these appointments have yet to be confirmed by the Senate. Mr. Batcheller is to succeed O. H. Caldwell from the first zone, and Professor Jansky is to succeed Sam Pickard from the fourth zone.

Mr. Batcheller was appointed to membership on the Board of Direction in 1928, and was elected to Board membership by the Institute for a three-year term beginning with 1929.

Professor Jansky was elected to membership on the Board of Direction by the Institute in the last election.

The Senate Committee has confirmed the reappointment of Judge Ira E. Robinson of the second zone, Judge Sykes of the third zone, and Mr. H. A. Lafount of the fifth zone.

Committee Work

1929 Committee Chairmen

At the February 6th meeting of the Board of Direction the Board approved the appointment of committee personnel for 1929, the chairmen of the various committees being as follows:

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Board of Editors, W. G. Cady; Committee on Nominations, Melville Eastham; Committee on Broadcasting, L. M. Hull; Committee on Sections, E. R. Shute; Committee on Admissions, R. A. Heissing; Committee on Constitution and Laws, R. H. Marriott; Committee on Membership, I. S. Coggeshall; Committee on Standardization, J. H. Dellinger; Committee on Meetings and Papers, K. S. Van Dyke; Committee on Publicity, W. G. H. Finch; Committee on Institute Awards, Melville Eastham.

Committee on Admissions

A meeting of the Committee on Admissions was held on February 6th in the office of the Institute in New York City. The following members were present: R. A. Heising, chairman; H. F. Dart, George Lewis, C. M. Jansky, Jr., and E. R. Shute.

The committee considered six applications for transfer or election to the higher grades of membership in the Institute.

Institute Meetings

Proposed Lehigh Valley Section

Much interest has been expressed on the part of the Institute members residing in the vicinity of Allentown, Pa., in the organization of a Lehigh Valley section of the Institute. B. H. Eckert and F. J. Hardner have been instrumental in the preliminary organization plans which brought forth a meeting of the Institute members at Muhlenberg College, Allentown, Pa. on February 8th, 1929. At this organization meeting B. H. Eckert and C. F. Maylott presided.

The paper by R. L. Smith-Rose, "Radio Direction-Finding by Transmission and Reception," was presented by Mr. Maylott. F. J. Hardner presented a paper, "Radio Noise Finding and Its Elimination." Messrs. Maylott, Hardner, Thomas, Muthard, and Kleck participated in the discussion which followed the presentation of these papers.

Mr. Maylott was elected chairman of the temporary organization. Another meeting is to be held on the 8th of March.

New York Meeting

At the New York meeting of the Institute held on February 6th in the Engineering Societies Building, 33 West 39th Street, A. Hoyt Taylor was inaugurated as President of the Institute. Ralph Bown opened the meeting with a short speech of introduction to Dr. Taylor.

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A paper by R. L. Smith-Rose, of the National Physical Laboratory, Teddington, England, on "Radio Direction-Finding by Transmission and Reception" was presented by L. M. Hull, of the Radio Frequency Laboratories, Boonton, New Jersey. Following the presentation of the paper the following participated in its discussion: A. Hoyt Taylor, L. M. Hull, Lester Jones, S. W. Dean, H. E. Hallborg, Stuart Ballantine, and others.

Three hundred and twenty-five members of the Institute and guests attended this meeting.

The April New York meeting, which is to be held on April 3rd, will consist of a symposium on frequency measurements. There will be seven contributions to this symposium by engineers intimately associated with frequency measurements and standardization.

Buffalo-Niagara Section

L. Grant Hector, of the University of Buffalo, presented a paper, "Apparent Equality of Loudspeaker Output at Various Frequencies", at the January 17th meeting of the Buffalo-Niagara Section. L. C. F. Horle, chairman, presided.

Messrs. Horle, Lidbury, Henderson, Stone, and others participated in a discussion which followed. Twenty-five members of the section attended the meeting. This paper is printed elsewhere in this issue of the Proceedings.

Cleveland Section

On January 25th a meeting of the Cleveland section was held jointly with the Cleveland Astronomical Society in the Case School of Applied Science, Cleveland, Ohio, presided over by Chairman Bruce W. David and attended by one hundred and sixty members and guests.

A paper was presented by H. T. Stetson entitled "Sunspots, Radio, and Weather." The paper showed the striking relations between solar activity and certain of the earth's meteorological conditions. Radio reception, the earth's magnetic field, and to a certain extent the weather are affected by sunspots. The paper was illustrated with lantern slides.

Los Angeles Section

The Los Angeles section held a meeting in the Elite Cafe, Los Angeles, on January 21st, attended by one hundred and twenty-five members and guests. T. F. McDonough, chairman of the section, presided.

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A paper entitled "The Nature of Sound, Its Transmission and Reproduction" was presented by A. P. Hill, of the Southern California Telephone Company.

Rochester Section

The Rochester section held a meeting in the Sagamore Hotel, Rochester, N. Y., January 11th, presided over by E. C. Karker, vice-chairman of the section.

Joseph P. Maxfield, of the Victor Talking Machine Co., presented a paper on "Physical Requirements of High Quality Audio-Frequency Reproduction." By means of lantern slides and a special orthophonic victrola, the speaker demonstrated improvements that had been made recently in the art of sound reproduction. The limitations of mechanical and electrical reproduction of sound were pointed out.

This was a joint meeting with the Rochester section of the American Institute of Electrical Engineers and the Rochester Engineering Society.

San Francisco Section

A meeting of the San Francisco section was held in the Hotel Bellevue, San Francisco, January 23rd, attended by twenty-eight members. Leonard F. Fuller, chairman, and Donald K. Lippincott, vice-chairman, presided.

Harry R. Lubcke presented a paper on "Vacuum-Tube Voltmeter Design." The election of officers was held with the following results: Donald K. Lippincott, chairman; Walter D. Kellogg, vice-chairman; Paul R. Fenner, secretary-treasurer.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

GEOGRAPHICAL LOCATION OF MEMBERS ELECTED FEBRUARY 6, 1929

Transferred to the Member grade

Ohio	Cleveland, WJAY, Schofield Bldg	Gray, Harold E.

Elected to the Member grade

Dist. of Columbia	Washington, 406 International Bldg	Stewart, Jessica Dee
Nicaragua	Managua U.S.S. Rochester, Balboa, C. Z.	Coulter, Howard N.

Elected to the Associate grade

Alabama	Mobile, c.o Reynolds Music House Co	Helt, Sanford
	Mobile, P. O. Box 801	Helt, Scott
California	Oakland, 1208 E. 18th St	Brearty, Lawrence S.
	Los Angeles, 444 Kings Road	Grosjean, G. M.
	San Francisco, 2370 Filbert St	Smith, Herbert H.
	San Jose, 398 N. 8th St	Kelliher, John Edward
Colorado	Colorado Springs, 920 E. Monument	Harvey, Francis M.
Dist. of Columbia	Washington, Bureau of Standards	Brattain, Walter H.
	Washington, 3706-24th St., N. W	Day, John F.
	Washington, 401-23rd St., N. W	Detwiler, Donald J.
Illinois	Chicago, 4134 North Richmond Ave	Brose, Fred O.
	Chicago, 2787 Frances Place	Edwards, Charles
	Chicago, 6257 Harper Ave	Hermanutz, Ray J.
	Chicago, 2251 E. 71st St	Timmings, George H.
	Oak Park, P. O. Box 552	Hoffman, Walter Henry
Iowa	Blairsburg	Baughman, Raymond C.
Kansas	Atchison, 624 U St	Gier, Willard Marion
Massachusetts	Arlington, 68 Marathon St	Dempsey, John P.
	Boston, P. O. Box 1678	Decker, Donald Philips
	Cambridge, 15 Holworthy Hall	Hunkins, Harold R.
	Cambridge, 112 Lakeview Ave	Noyes, Atherton, Jr.
	Cambridge, Harvard University, Holworthy 16	Packard, Alden C.
	Cambridge, 15 Holworthy Hall	Taylor, John Pratt
	Medford, 21 Frederick Ave	Surette, Dennis C.
	N. Wilbraham, Cottage Ave	Garvey, Edmond
	Springfield, 407 St. James Ave	Chapman, Alan B.
	Springfield, 19 Elliott St	Van Doeren, C. A.
Michigan	Detroit, 11338 Dexter Blvd	Huber, John E. L'E.
	Detroit, 504 Commerce Bldg	Kratokvil, Frank M.
	Detroit, 15364 Oakfield Ave	Martin, Robert D.
	Detroit, 16508 Welland Ave	O'Neil, John J.
	Detroit, 135 Nevada, W	Porter, Samuel
	Jackson, 1019 First St	Rich, C. E.
	Jackson, 434 Stewart Ave	Riethmiller, Earl R.
Missouri	St. Louis, 6339 Marquette Ave	Riddle, Ruston L.
Nebraska	York, 1214 Lincoln Ave	Meyer, Albert
New Jersey	Allendale, P. O. Box 207	Asten, Oliver B.
	Atlantic City, 225 Atlantic Ave , Apt. B-6	Enderle, Jackson J.
	Irvington, 617 Grove St	Hart, James J.
	Linden, 1104 Wood Ave	Holetz, Alexander C.
New York	Brooklyn, 54 Wyckoff St	Kirdahy, Emil
	Buffalo, 15 University Ave	Schwing, Russell L.
	Great Neck, 20 Arrandale Ave	Genner, W. Gordon, Jr.
	New York City, 336 East 5th St	Barnueba, Richard
	New York City, 116 Broad St	Danz, Hermann
	New York City, 273 West 113th St	Diaz, Ernest
	New York City, 120 East 30th St	Hastings, Gerald M.
	New York City, Elec. Research Products, Inc., 250 W. 57th	Hipp, W. S., Jr.
	New York City, 463 West St., Room 279	Kurtinaitis, John V.
	New York City, 152 Dyckman St	Murphy, E. Edward
	New York City, Elec. Research Products, Inc., 250 W. 57th	Padden, Cecil John
	New York City, 269 W. 34th St	Payette, Walter S.
	Riverhead, L. I., c/o Radio Corporation of America	Henery, R. S.
	Schenectady, 221 Seward Place	Biver, Carl J.
	Schenectady, 2 Eagle St	Frink, Frederick W.

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	Schenectady, 607 Chapel St.	Lynn, Roland A.
	Schenectady, Y.M.C.A. Bldg.	McLennan, Miles Ayrault
	Schenectady, 614 Campbell Ave.	Richards, Philip A.
	Schenectady, Y.M.C.A., Room 425	Tanke, Harold F.
North Carolina	Norlina, Norlina Hotel	Schulke, Enno
Ohio	Akron, 179 Ido Ave.	Smith, Paul C.
	Cleveland, 4124 Bailey Ave.	Deutsch, E. J.
	Cleveland, 3905 Svoboda Ave.	Pitonak, Joseph C.
	Cleveland, 2845 Prospect Ave.	Umbrecht, Stanton
	Lakewood, 1288 Ramona Ave.	Drummond, P.
	Marion, 360 Silver St.	Ackerman, Francis R.
	Newphiladelphia, 334 Minnich Ave.	Nolan, John
	Salem, 604 McKinley Ave.	Ludlow, Gilbert H.
	Warren, 1903 Grove Place	Wason, R. E.
Oklahoma	Tulsa, Radio Dept., Skelly Oil Co.	Rand, George L.
	Tulsa, c/o Dodge Electric Co., 318 So. Boulder	Swan, Merrill LeRoy
Pennsylvania	Allentown, 959 Turner St.	Keck, Kenneth K.
	Boyertown, 36 North Reading Ave.	Markle, J. E.
	Philadelphia, 4916 Chestnut St.	Kienzle, D. R.
	Wayne, 225 W. Wayne Ave.	Adelberger, Paul J.
South Dakota	Huron, Box 663	Griffiths, A. Elmer
West Virginia	Kingwood, 134 Price St.	Michelson, Peter
Tennessee	Saltillow	Berry, James S.
Texas	Houston, 403 Studewood	Harold, George
Virginia	Quantico, Post Radio Station, Brown Field	Beardsley, Franklin
Washington	Seattle, 4055-42nd Ave., S. W.	Mood, George T.
Wisconsin	Fond du lac, 284 Dixie St.	Eckert, Walter R.
	Madison, 1530 Jenifer St.	Hagen, Paul A.
	Milwaukee, 672 Grove St.	Frohrieb, Edward F.
Canada	Hamilton, Ont., 339 Wilson St.	Clemnts, B. D.
	Hamilton, Ont., 267 Cannon St., E.	Crofte, Cecil T.
	Hamilton, Ont., 65 Garfield Ave S.	Bain, J. R.
	Penticton, B. C.	Howse, J.
	Toronto, Ont., 67 McLean Blvd.	Clarkson, Percy W.
	Toronto, Ont., 578 Spadina Ave.	Freeman, William
	Toronto, Ont., 41 McPherson Ave.	Glassford, J. O.
	Toronto 12, Ont., 1331 Avenue Road	Kinnear, Donald R.
	Toronto, Ont., 578 Spadina Ave.	Southam, Sheraton
	Toronto, Ont., Ontario Forestry Branch, Parliament Bldg.	Ward, Charles
Channel Islands	Guernsey, 16 Bordage St.	Laker, Edward W.
China	Canton, Sun Yatsen University	Chu, Chih Teih
Denmark	Copenhagen, Vennemindevej 3, 3 Str.	Svenningsen, Karl
England	Branston, Lincoln	Sparrow, A. W.
	Cambridge, Corpus Christi College	Landale, S. E. A.
	London, Berkeley St., c/o Thomas Cook, Ltd.	Samarasekara, V. R.
	London, W1, Berkeley St., c/o Thomas Cook, Ltd.	Wijeyeratne, P. de S.
	Swindon, Wilts, 32 Prospect Hill	Mustchin, Norman
	Whitley Bay, Northumberland, 25 Marden Road South	Gow, Charles N.
Germany	Berlin, Wilmersdorf, Hildegardstr. 13 b	Lock, M. J.
	Heilbronn, Bockingen	Hertweck, C.
New Zealand	Dunedin, 418 Anderson's Bay Road	McKewen, J. D.

Elected to the Junior grade

California	Los Angeles, Y.M.C.A., 715 Hope St.	Belleville, Logan
Colorado	Denver, 1237 Elizabeth St.	Foraker, William Nelson
Illinois	Chicago, 5211 Kimbark Ave.	McCammon, Donald
	Paw Paw	Braffet, Donald H.
Kansas	Coffeyville, P. O. Box 190	Russell, Charles
New York	Buffalo, 17 William St.	Freedman, Sam
	Schenectady, Y.M.C.A., State St.	Lee, Samuel T., Jr.
Ohio	Lakewood, 12974 Emerson Ave.	Thompson, J. Kent
Pennsylvania	Philadelphia, Mt. Airy, Thouron Ave. and Gorgas Lane	Ellinwood, Kenneth M.
England	London N 5, 59 Highbury, New Park	O'Rourke, Sydney P.

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APPLICATIONS FOR MEMBERSHIP

Applications for election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to election of any of these applicants should communicate with the Secretary on or before March 29, 1929. These applicants will be considered by the Board of Direction at its April 3rd meeting.

For Election to the Associate grade

California	Los Angeles, c/o California Radio Service, 800 N. Spring St.	Herrnfeld, Frank P.
	Los Angeles, 5308 Third Ave.	Howe, Franklin Joachino
	Los Angeles, 521 Amethyst St.	Oodrys, Arthur
	Palo Alto, Federal Telegraph Co.	Shermund, Ralph C.
	San Jose, 1148 Lincoln Ave.	Saine, Harry
	Stanford University, Box 1331	Skinner, Clifton Ross
Connecticut	Bridgeport, 68 Mead St.	Tillman, John Elmer
	Kensington, P.O. Box 276	Edgerly, A. H., Jr.
Dist. of Columbia	Washington, Radio Section, Bureau of Standards	Arnold, Prescott N.
	Washington, 1705 Bay St., S.E.	Bower, Ward E.
	Bellevue, Naval Research Laboratory	Hyland, Lawrence A.
	Washington, 2301 Cathedral Ave., N.W.	Jackson, William E.
	Washington, 900 F St., N.W.	Kaplan, Jacob
	Washington, 2618-13th St. N.W.	Moulton, Theodore S.
	Washington, The Chastleton, 1701 16th St. N.W.	Newton, Jane Elizabeth
Illinois	Chicago, 5054 Agatite Ave.	Beusman, Robert Marshall
	Chicago, 6716 Parnell Ave.	Cubert, Joseph R.
	Chicago, 411 S. Ashland Blvd.	Kempf, Frank J.
	Chicago, 3250 North Crawford Ave.	Kincaid, Owen D.
	Chicago, 501 North Central Ave.	Knutson, Ben M.
	Chicago, 1706 N. Mayfield Ave.	Lamb, Harold A.
	Chicago, 3823 S. Kedzie Ave.	Patterson, Joseph A.
	Chicago, 3621 E. 107th St.	Siovic, John E.
	Chicago, 7728 Calumet Ave.	Van Sickel, R. E.
	Lake Forest, 1199 Edgewood Road	Baker, James Maurice
	Litchfield, Box 214	Hopper, C. L.
Indiana	Evansville, 112 Jackson Ave.	Norwood, Vandle Clarence
Iowa	Des Moines, 1538-31st St.	Hutchison, Sam T.
Kansas	Pittsburg, Pittsburg Radio Service	Forman, H. W., Jr.
Louisiana	New Orleans, 2625 Jefferson Ave.	Goldstein, Henry R.
	New Orleans, 4614 Carondelet St.	Adler, Leonard E.
Maine	Yarmouth, Greeley Road	Tamminen, Nestor
Massachusetts	Belmont, 98 Payson Road	Anthony, A. W., Jr.
	Boston, 472 Massachusetts Ave.	Nunez, F. J.
	Boston, Custom House, c/o Supervisor of Radio	Weston, Irving L.
	Cambridge, 22 Bigelow St.	Battison, Wallace A.
	Cambridge, General Radio Co., 30 State St.	Thiessen, Arthur E.
	Chatham, c/o Radio Corporation of America	Thing, K.
	North Attleboro, 9 Elm St.	Hale, Willis L.
	Revere, 62 Malden St.	Fletcher, Frederick R.
	Salem, 28½ Grove St.	Heffernan, S. K.
	Springfield, 62 Kimberly Ave.	Ferguson, George Well
	Stockbridge, Box 982	Harrington, George N.
Michigan	Alpena, 111 Tawas St.	Snyder, Leslie T.
	Ann Arbor, 1030 Church St.	Mathison, Gerald
	Detroit, 3028 Lothrop Ave.	Woehr, William
Minnesota	Minneapolis, 1521 University Ave., S. E.	Reed, Harry F.
Missouri	Bucklin	Holmlund, A. Earle
	Mendota, R.F.D.	Davis, Glen A.
	St. Louis, 4047 W. Pine St.	Humphreys, Irl W.
Nebraska	Falls City, 1801 Morton St.	Chesley, Arthur D.
New Hampshire	Claremont, 227 Main St.	Hodge, V. W.
New Jersey	Bloomfield, 25 Grace St.	Henry, T. J.
	Camden, 303 North 6th St.	Galvin, Robert
	Clayton, 326 Broad St.	Purvis, Charles G.
	Deal, Box 122	Shaw, Robert C.
	Glen Ridge, 129 Midland Ave.	Grabo, Irving C.

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	Linden, 104 Luttgen Place	Angle, Gafford Brock
	New Brunswick, R.C.A. Radio Station	Bohman, Victor A.
	New Brunswick, 149 Hale St.	Lucas, Earle F.
	Stanhope, Box 261	Peterson, Arthur C.
New York	Astoria, L. I., 14-34 Grand Ave.	Ractliffe, Charles Lionel
	Brooklyn, 392 Sackman St.	Berner, Aaron
	Brooklyn, 1121 Bedford Ave.	Dean, Leon W.
	Brooklyn, 506 Amboy St.	Epstein, Reuben
	Brooklyn, 421 Chester St.	Haynes, Nat
	Brooklyn, 1300 New York Ave.	Herdman, Raymond C.
	Brooklyn, 1257 New York Ave.	Hesse, Henry Richard
	Brooklyn, 1859-62nd St.	Sass, Isidore
	Brooklyn, 55 Johnson St.	Skinker, Murray F.
	Buffalo, 191 Franklin St.	Bandetson, Harry
	Greenport, 829 Main St.	Larsen, Carl L.
	Jamestown, 72 Campbell Ave.	Beaumont, William Frederick
	New York City, 1050 Park Ave.	Cohn, Ralph I.
	New York City, 617 West 141st St.	Kahn, Morton B.
	New York City, 760 West End Ave.	Littman, Leon
	New York City, 160 West 100th St.	Lopez, Melchor, Jr.
	New York City, 2740 Marion Ave., Bronx	Myers, Theobald
	New York City, 500 Riverside Drive	Sicari, Domenic
	New York City, 404 Riverside Drive	Turner, Eugene T., Jr.
	New York City, 781 East 182nd St., Bronx	Wheeler, George D.
	Richmond Hill, 8512 110th St.	Hudtwalker, William Theodore
	Riverhead, P.O. Box 982	Trevor, Bertram
	Rochester, 55 Hortense St.	Dwyer, Vincent J.
	Rochester, 207 Ave. C	Wiebach, William T.
	Rocky Point, c/o R.C.A.	Goldstine, Hallan E.
	Schenectady, 105 Seward Place	Clarke, Varro J.
	Schenectady, 422 Y.M.C.A. Bldg.	Lynn, L. H.
	Schenectady, 812 Union St.	Orr, Robert W.
	Whitestone, L. I.	Summers, Llewelyn L. B.
Ohio	Akron, 160 Fir St.	Ehrisman, Henry O.
	Ashtabula, R.D. No. 2	Andrus, Roy E.
	Cincinnati, 1212 Sassafras St.	Verkley, B. M.
	Cincinnati, 3484 Vine St.	Wells, Martin M.
	Columbus, 34-18th Ave.	Auckerman, U. H.
	Columbus, Ohio State University, E.E. Dept.	Roetken, A. Allen
	Dayton, 43 Victor Ave.	Franzwa, Frederick J.
	Gambier, Kenyon College	Cottrell, Casper L.
	Lakewood, 2028 Waterbury Road	Hood, William A.
	Marion, 360 Silver St.	Ackerman, Francis R.
	Niles, 424 Allison St.	De Cola, Rinaldo
	Youngstown, 3630 Market St.	Pennock, P. L.
Oklahoma	Norman, 132 Page St.	Moffett, Le Roy, Jr.
	Oklahoma City, 1624 East Park Place	Pata, Yaromir J.
	Tulsa, Radio Station KVOO	Golder, Frank E.
Oregon	Portland, 605 Marguerite Ave. N.	Trumbull, A. F.
Pennsylvania	Allentown, 531 N. 7th St.	McGee, Michael J.
	Easton, 528 Centre St.	Messinger, Reuben R.
	Lancaster, 36 S. Lime St.	Russell, Walter L.
	Narberth, 403 N. Narberth Ave.	Bates, Clifford W.
	Philadelphia, 2030 E. Hazard St.	Martino, Alphonso E.
	South Tamaqua	Delp, Paul L.
	Upper Darby, 297 Springton Road	Lewis, Oliver I.
	Wilkinsburg, 901 South Ave.	Armstrong, Ralph W.
Rhode Island	Providence, 160 Cypress St.	Adams, Raymond R.
Tennessee	Memphis, 1961 Harbert St.	Brooks, Maurice W.
Texas	Dallas, 2603 Madera St.	Bennett, Porter T.
	Dallas, 1205½ Elm St.	Godard, L. G.
	Dallas, 5004 Goodwin Ave.	Melroy, Harry C.
	Fort Worth, 907 W. T. Waggoner Bldg.	Zeidlik, William J.
	Moody	McCauley, E. Ray, Jr.
Virginia	Alexandria, Box 107 R4	Meyers, Paul F.
	Marion	Cummings, G. N.
Washington	Seattle, 1833-13th Ave.	McAvoy, Edward J.
Wisconsin	Menasha, 526 Keyes St.	Peerenboom, Cyril A.
	Sheboygan, 922 Clara Ave.	Flentje, Le Roy G.
	Stoughton, 313 S. Academy St.	Turner, Russel S.
	Wausau, 108 Grand Ave.	Krueger, Otto J.
Canada	Montreal, Que., Northern Electric Co., Ltd., Keefer Bldg.	Cash, J. Allan
	Montreal, Que., 89 Laurier Ave. E.	Minore, J. L.
	Montreal, Que 4581 Sherbrook St. W.	Wilcox, B. B.
	St. Hyacinthe, Que, 7 Bourassa St.	Chagnon, Adolphe

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Denmark	Copenhagen, Kastelsvej No. 3	Bertzow, Johannes Andreas
England	Chelmsford, Essex, 26 Queen's Road	O'Neill, R. F.
	Leigh-on-Sea, Essex, 106 Western Road	Richardson, F. C.
	London, N.W.6, 42 Fairhazel Gardens, Hampstead	Erdman, H.
	London, SW5, 12 Trebovir Road	Megaw, E.
	London, Berkeley St., Thos. Cook & Son, Ltd.	Samarasckara, V. R.
	Parkstone, Dorset, St. Nicholas Castledene Road	White, John Walter
Holland	2 Dedelstraat den Haag	Bartelink, E. H. B.
Japan	Kumamoto, Kumamoto Broadcasting Station JOGK	Hirose, M.
	Kumamoto, Kumamoto Broadcasting Station	Morita, Kazuyoshi
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PART II

TECHNICAL PAPERS

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

RADIO DIRECTION-FINDING BY TRANSMISSION AND RECEPTION

(With Particular Reference to Its Application to Marine Navigation)*

By

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(The National Physical Laboratory, Teddington, England)

Summary—*This paper presents a critical résumé of the performance of apparatus employed for radio direction determination either by transmission or by reception. After an historical summary of results obtained in various parts of the world, a brief description is given of the fundamental principles underlying radio direction-finding. In this section attention is drawn to the application of the principle of reversibility to this art, by the aid of which the behavior of directive radio transmitters can be largely predicted from the more numerous results and greater experience already obtained with directional receivers.*

The next two sections of the paper give a review of the results obtained in Great Britain during the course of extensive investigations into this subject during the past seven years. Observations obtained from thirteen direction-finding receiving stations, specially erected for the purpose, have been carefully analysed and the performance of the apparatus studied under a variety of conditions, including operation in daylight and darkness, and both oversea and overland. In addition, some two years have been spent in studying the performance of a rotating-loop beacon transmitter, by means of which accurate radio bearing can be obtained with any type of receiving apparatus.

The later portions of the paper deal with the application of direction-finding to marine navigation, and of the possible effect of coastal and night errors in connection therewith. The production of night errors on closed loop receivers by the horizontal component of the electric force in downcoming waves is explained, and a demonstration is given of the manner in which the Adcock aerial system gives freedom from such errors. The paper concludes with a discussion of the relative advantages of direction-finding by transmission and reception for navigation purposes. A bibliography of the subject is appended.

* Dewey decimal classification: R125.1. Original manuscript received by the Institute, November 14, 1928. An abstract of this paper was read before the meeting of the Union Radio-Scientifique Internationale held in Brussels, September 5 to 10, 1928. Presented before New York meeting of the Institute, February 6, 1929.

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1. Historical

THE application of the radio direction-finder both as a navigational instrument and as a useful scientific tool in the study of the propagation of electromagnetic waves has been developing rapidly during the past few years. As a result, a number of text-books expressly devoted to the subject of directional wireless have appeared, and the reader must be referred to these[1]† and other published works for an account of the fundamental principles of the art of direction-finding and its historical development. The object of this paper is to present a critical resumé of the results of investigations which have been carried out over a period of several years into the performance of direction-finding systems, and the historical background provided in this section will be confined to this aspect of the subject.

(a) *The Direction-Finding Receiver*

The modern radio direction-finder undoubtedly owes its success very largely to the introduction of valve amplifiers, enabling a moderately large reception range to be obtained, and its practical development therefore dates from about 1915. Previous to this, such systems of direction-finding as were in existence were confined to short-distance working and the comparatively crude instruments then in use made accurate systematic observations difficult to obtain. As early as 1908, however, Pickard[2] observed that large errors might be obtained in the reading of coil direction-finders due to buildings, trees, and other obstacles in the neighborhood. In the diagrammatic representation of his results, the errors are shown to be approaching 90 deg. It was found also by Fessenden[3] in the years 1901–07, that errors in apparent direction of as much as 20 deg. to 45 deg. might be obtained in the indication of these instruments when receiving over a range of 100 miles. These errors were attributed to a refraction effect resulting from the difference in conductivity of land and sea-water, or even to a varying local conductivity of the ground and of vegetation. In making continuous observations day and night for a week, Fessenden was apparently the first to observe that the errors were greatest during the night, a fact which he attributed to a refraction effect of large clouds of ionized air in the path of the waves.

† See bibliography.

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A method of obtaining an absolute zero of signal-strength on a small frame-coil direction-finder was described by A. H. Taylor[4], in which the "antenna effect" of the coil is compensated for by a small emf from an auxiliary frame at right angles to the first. Using this system at Washington, it was found that while readings taken in the daytime were fairly accurate those taken near sunset and at night were very erratic. While it appeared that the variations observed on continuous waves of wavelength 13,600 meters were greater than on shorter waves, they were quite serious on damped waves of 1,500 meters wavelength. These variable results were briefly ascribed to reflection and refraction effects occurring during the propagation of the electromagnetic waves over the earth's surface.

The liability of the metalwork of a ship to produce a quadrantal error in the readings of a direction-finder mounted thereon was mentioned by Blondel[5] in 1919, while the corresponding effects on an aeroplane were later described by Robinson.[6] The complete theory of the effect of the metal hull of a ship on a direction-finder was first given by Mesny[7] in 1920. Calculations from this theory, confirmed by experimental results, showed that the quadrantal error obtained may be as great as 12 deg. This error was shown to be independent of wavelength and to decrease with the height of the frame coil above the deck. Attention was also drawn in the above paper to the approximately analogous case of a direction-finder erected upon a hill or an island, and the resulting quadrantal error which may be obtained in such a case is illustrated by a curve having a maximum value of 15 deg.

The phenomenon of the refraction of electromagnetic waves in passing over a surface of suddenly changing conductivity was discussed by T. L. Eckersley[8] in 1920. Experimental observations made in Cyprus and Egypt on wavelengths between 800 and 1,100 meters showed that wireless waves in crossing a coastal boundary between sea and land might suffer a deviation of as much as 4 deg. This deviation falls to zero for normal incidence of the waves on the coast line, and it was also shown to be negligible for wavelengths exceeding 2,000 meters. In the paper there is quoted an interesting case of a bad day minimum being produced by the reception of two waves from a transmitter, the two waves arriving by different paths and with a phase difference which resulted in a rotating field.

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Somewhat similar refraction effects on wireless waves passing from dry to wet ground and across a river were mentioned by Kiebitz[9] in connection with experiments on a directional transmitter. The deviations of the waves amounted to 8 deg. or 9 deg. for a wavelength of 550 meters.

Some results showing the errors to which a radio direction-finder may be subject due to local conditions were given by Hollingworth and Hoyle.[10] Masses of metalwork, tuned circuits, and overhead wires were found to produce appreciable errors in the readings.

In a most valuable paper published in 1920, Round[11] gave an account, chiefly from his personal experience, of the development and application of the direction-finder during the war. The manner in which the instrument was perfected as a useful tool for both military and naval purposes was described together with the various types of errors encountered, both by day and by night. Reference was made to the work of Adcock, Eckersley, and Wright in connection with these errors, and a brief indication was given of the means by which they might be eliminated for practical direction-finding purposes.

A large amount of experimental work on the intensity and directional properties of the electro-magnetic field radiated from an aeroplane transmitter was described by Baldus, Buchwald, and Hase in 1920,[12] while a mathematical treatment of this case was given by Burstyn.[13] The errors in the apparent bearings of an aeroplane at a ground direction-finding station were discussed in detail in their relation to the plane of polarization of the emitted waves. The experiments of Baldus and Buchwald, in particular, showed that a closed-coil direction-finder on the ground could give errors of as much as 60 deg. in the bearings of aircraft. This fact is interesting in connection with the patent filed by Adcock[14] in 1919, in which was described a means of eliminating the error of observation of the orientation of aeroplanes.

Some further observations on the variable night errors were published in 1920 by Kinsley and Soby.[15] Variations in the apparent bearings ranging up to 50 deg. were recorded on wavelengths between 960 and 17,300 meters, and for ranges of transmission from 40 up to 7,500 miles.

Towards the end of 1920, the Department of Scientific and Industrial Research in England, acting on the advice of its

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Radio Research Board, began the erection of a number of directional radio receiving sets in various parts of the British Isles, attached either to Universities or to Government experimental establishments. The object of these installations was to make regular observations of the apparent radio bearings of various transmitting stations in order to obtain data on the nature, magnitude, and other characteristics of the variations of bearings which were previously known to take place. The general organization and conduct of this investigation was carried out by the author from The National Physical Laboratory, Teddington, as headquarters. Up to the time of the termination of the general investigation in September, 1926, about a quarter of a million observations had been accepted for correlation, these observations covering the range of wavelengths of 300 to 20,000 meters. A detailed record of the work, with discussion of the conclusions drawn therefrom, has formed the subject of a number of official reports published by H. M. Stationery Office, England.[16] In addition to these publications, descriptions of various investigations subsidiary to the main line of research have been published elsewhere. For example, in order to obtain accurate bearings in continuous wave working it was found necessary to prevent any mutual induction between the local oscillator and the receiving frames of a Robinson direction-finder by suitably screening the local oscillator.[17] Such screened oscillators have now been in use for several years in connection with various direction-finding work, and nowadays they find considerable application in radio measurement work.

These experiments were later extended to that of screening assemblies of amplifying and receiving apparatus to prevent the undesired direct induction of signals. Data were obtained by Barfield[18] for the screening of a complete hut containing the receiving apparatus as, for example, in the Bellini-Tosi direction-finding system. In the course of the same investigation Barfield demonstrated the properties of open wire screens, and these were later applied to the reduction of the error known as "antenna-effect" in radio direction-finding.

Various experiments were carried out in the early stages of the main investigation to ascertain the effect of local conditions such as metal-work, overhead wires, trees, etc., on the readings of direction-finders.[19,20] These experiments showed that some quite large errors, ranging up to 22 deg., can be produced by the

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proximity of such obstacles and emphasize the importance of exercising care in selecting a suitable site for a direction-finding installation. The difficulty in finding any approach to an ideal site was illustrated by the results obtained from the stations selected. In very few cases was the error due to local conditions less than 2 deg. It is fortunate, however, that such a type of error remains constant in value for any particular direction, so that it can be treated as of the nature of a permanent deviation, which can be ascertained periodically from a calibration of the station. In the case of the Aberdeen station, the cause of permanent errors ranging up to 15 deg. was traced to a long iron plate beneath the ground and supporting a sewer duct, over which the direction-finding installation was inadvertently erected.[20]

During the course of the main investigation each of the three practical types of direction-finder, known as the Bellini-Tosi, Robinson, and single-coil systems, respectively, has been used in some portion of the investigation: and it was naturally desirable to verify whether comparable results could be obtained with any of these systems. A simple consideration of the theory shows that there is no essential difference in the basic principle of any of the systems, for in each case the reading of the direction is made as the result of some portion of the apparatus rotating about a vertical axis being set in the position in which there is a minimum, or in the ideal case, zero emf induced by the incoming waves. This theory was discussed in a previous publication[21] which also contained a description of the results of experiments showing that each system was equally subject to variable night errors.

(*b*) *The Direction-Finding Transmitter*

As an alternative to the use of a special radio receiving instrument for the determination of bearings the directional property can be transferred to the transmitting station, so that its direction can be determined at a distant receiving station by the aid of some characteristic of the emitted radiation.

Both of the types of directional transmitting systems which are in use today may be said to date from the time of Hertz, since in his experimental researches Hertz used reflectors to concentrate the radiation from a straight rod aerial and also loops whose transmission or reception properties depend upon the orientation of the loop. Of recent years the properties of reflecting

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systems used in conjunction with linear aerials have been studied by Marconi,[22] Franklin,[23] and others,[24] and with the application of the beam system to long distance communication considerable development has been taking place in many countries. The beam system has also been adapted as a rotating beacon transmitter,[25] but practical limitations necessitate its operation on the very low wavelengths of 6 to 10 meters. Experimental beacons of this type have been installed at Inchkeith[22,23] and South Foreland,[25] but it is believed that its use among ships has not been very widespread.

The predecessor of the closed loop antenna, whether for reception or transmission, is to be found in either the inverted *L* aerial or in a pair of spaced vertical aerials. The former has developed into the Beverage aerial which is now used as one method of obtaining directional selectivity at the receiving end. A system of inverted *L* aerials giving directional transmission has been experimented with by Scheller, Buchwald,[26] and Kiebitz.[27] For navigation purposes it is also probable that the Telefunken compass[28] arrangement falls under this heading. This system made use of a series of directional aerials radiating from a central mast, these arials being excited in turn by a rotary switch operating at a speed of one revolution per minute. A non-directional aerial was also provided for the purpose of transmitting a timing signal applicable to the directive system. By noting the interval between this time signal and the reception of the signal from the directional system, the bearing of the transmitter could be estimated from a distant receiver.

The combination of the pair of spaced aerials developed first into the Bellini-Tosi system of direction-finding and later into the rotating frame coil. The transmitting equivalent of the Bellini-Tosi direction-finder is to be found in the Radiophare, several of which were in operation on the French coast prior to 1914. In the same manner the single-frame coil can be adapted to transmission since its polar radiation characteristic is the well-known figure-of-eight diagram, and thus the intensity of the radiation varies according to a cosine law with the angle between the plane of the loop and the direction of transmission. One of the great advantages of a loop transmitter is that it can be operated on the wavelengths usually employed in ship and aircraft wireless communication. Various methods of applying the rotating loop transmitter for navigation purposes were

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described by Erskine-Murray and Robinson in 1922,[29] while more recently Gill and Hecht[30] have recounted the development of the rotating beacon by the British Air Ministry and supplied some typical results obtained in the application of this system to the navigation of aircraft.

The use of two directional transmitting aerials sending complementary Morse signals (such as *A* and *N*) upon the same wavelength for the purpose of providing an equi-signal zone along a fixed course appears to have originated with Scheller in Germany, and to have been investigated by Buchwald[26] and Kiebitz[27] in 1920. With the substitution of closed loops for the open inverted *L* aerials previously employed, Engel and Dunmore[31] published in 1924 an account of experiments designed to illustrate the utility of this system for the navigation of both ships and aircraft. This method is not strictly one of direction-finding since the observer or navigator is not generally able to determine his bearings or position by this means, but only to locate himself along a fixed course as given by the beacon transmitter. For this reason the system is more directly applicable to the navigation of aircraft flying along fixed routes than to the navigation of ships; and it is from this point of view that considerable development of the method has recently been carried out in the United States of America. This work has been described by Dellinger and Pratt,[32] who have applied a modified form of the radiogoniometer to enable the course or equi-signal zone given by the transmitter to be oriented in any required direction. The limitation of range of the system for accurate direction indications is illustrated by the results of some experiments reported by Pratt[33] in a separate paper. In common with all closed loop transmitters and receivers for direction-finding this system becomes liable to serious errors at ranges exceeding about 100 miles, even when the receiver is some 2,000 feet above the ground.

For further detailed information on the application of radio communication to aircraft navigation, the reader may be referred to the Bibliography recently published by Jolliffe and Zandonini.[34]

2. Brief Description of Methods of Direction Determination by Radio Transmission or Reception

(a) *The Radio Direction-Finder*

The radio direction-finder is now well known as an instrument which can be used to determine the direction of arrival of

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wireless waves, and it will suffice to give the briefest outlin heree of the principles underlying this instrument. The several commercial types of direction-finder now in use employ the same fundamental principle of the reception of vertically polarized* wireless waves by a frame coil. In Fig. 1(a), for example, let *C* represent a plane vertical loop rotating about a vertical axis in the field of an arriving wave whose component electric and magnetic forces are as shown. From the plan view in Fig. 1(b)

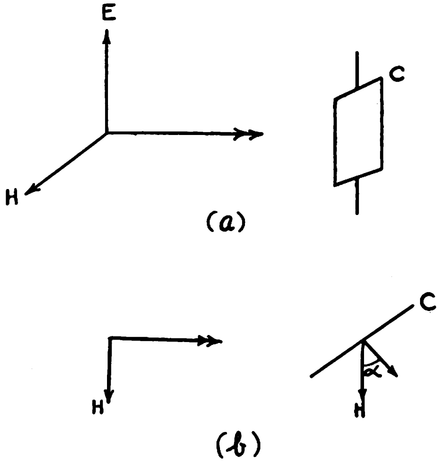

Fig. 1

it is evident that the emf induced in the loop by the arriving waves will be proportional to the cosine of the angle a between the direction of the magnetic field and the axis of the coil. The accuracy with which any definite position of the coil may be located depends upon the rate of change of emf with orientation, i.e., the accuracy is proportional to sin a. Thus the determination of the direction of arrival of the waves is most accurate when the signal emf induced by the waves is zero. This point is also illustrated by Fig. 2, which shows the theoretical polar reception diagram for a rotating loop. The strength of the signal emf induced in the loop by a wave arriving from any direction is proportional to the intercept of the vector *OA* made by the "figure of-eight."

The most important feature which requires attention in the design of a practical direction-finder is the avoidance of spurious

* The term "vertically polarized" is used here to indicate that the electric force of the wave lies in the vertical plane of propagation of the wave.

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emf's introduced into the system from one or both of the phenomena commonly known as "antenna-effect" and "direct pick-up." The term "antenna-effect" is applied to the property possessed by a frame coil receiver of acting as an untuned vertical aerial as well as a coil for reception purposes. As a result of this, the receiving system may have induced in it an emf whose phase and magnitude are independent of the orientation of the coil. The signal heard in the telephones will be the sum of that produced by the rotating coil as such and that due to the equivalent aerial effect of the whole receiver. As the coil is rotated it is found that the signal zeros become blurred into broad minima only, and moreover, they may be displaced from their correct positions. The existence of this antenna effect in the system

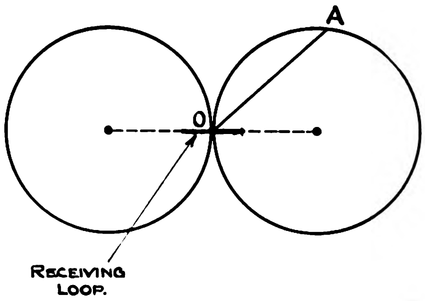

Fig. 2

therefore makes the observed directions incorrect, and also makes the determination of these directions more difficult.

Somewhat similar results may be produced by the second of the two causes mentioned above, viz. "direct pick-up." This last term implies that portions of the receiving system, such as the tuning circuits and the amplifier, are having emf's induced in them directly by the incoming waves. These emf's will obviously be independent of the orientation of the main receiving frame, and they will be effective in adding to or subtracting from the signal strength finally heard in the telephones. It must be appreciated that while these stray emf's may be small compared with the main emf picked up by the rotating frame-coil in its maximum position, they become of very great importance when the coil is turned into its minimum position.

The methods adopted for overcoming the effects of these spurious emf's are based on the use of somewhat elaborate

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screening arrangements, with or without the addition of a compensating condenser for the antenna effect. The application of these methods to the practical arrangement of single-coil direction-finders has been described by Kolster,[35] Dunmore,[36] Long,[1(g)] Mesny,[(f)] and the author;[37] and the reader must be referred to these descriptions for further details of such apparatus.

With the object of enabling the incoming signal to be clearly audible throughout the whole process of taking a bearing, a crossed-coil arrangement was described by Robinson[38] in 1920. This system was, at one time, particularly favored for use on aircraft, but with some modification described by Bainbridge-Bell,[39] it now finds widespread application to marine navigation.

Before the development of valve amplifiers made possible the use of rotatable multi-turn loops, Artom in 1903 and Bellini and Tosi in 1907 suggested and used large frames of a triangular shape, with the ends open at the top apex, for directional wireless communication. The arrangement developed from this and now generally known as the Bellini-Tosi system was fully described in 1908.[40] The large, fixed closed loops employed in this system are connected to a radiogoniometer, an instrument which reproduces in miniature the directive properties of the external field of the waves. Recent developments of this system for use in ships have resulted in the employment of smaller multi-turn loops, which are more conveniently fixed on board. The development and use of the Bellini-Tosi system during the war was described by Round[11] in 1920, while Horton,[41] Slee,[42] and Mesny[43] have dealt with some more recent developments.

(*b*) *The Rotating Loop Transmitter*

As an alternative to the direction-finding schemes outlined above, the directional part of the wireless system may be transferred from the receiving to the transmitting end. This is effected in the rotating loop beacon system, which has been developed to a high degree in Great Britain by the Royal Air Force, and which employs a vertical closed loop transmitter arranged to rotate about a vertical axis at a uniform speed of one revolution per minute. The polar radiation diagram of such a loop will be of the same figure-of-eight form as that shown in Fig. 2 for a loop receiver. Thus as the loop rotates the field radiated in any given direction will vary according to a cosine

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law, passing through successive maximum and minimum values at intervals of fifteen seconds. When the plane of the coil is perpendicular to the geographical meridian a characteristic signal is emitted by the beacon which may be termed the *N* point. An observer at a distant receiving station upon hearing this signal starts a chronograph. As the beacon rotates the intensity of the received signal varies and will ultimately pass through a minimum or zero value, at which instant it is known that the plane of the transmitting loop is at right angles to the great circle joining transmitter and receiver. If the reading of

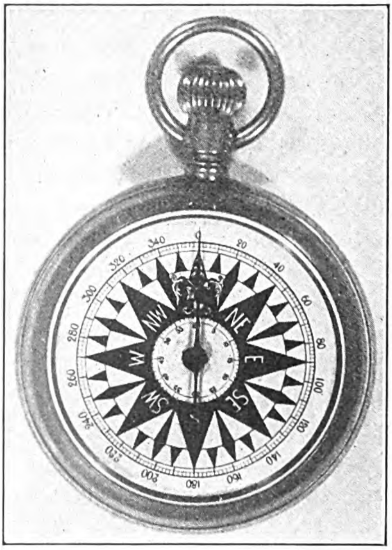

Fig. 3—Stop Watch Fitted with Special "Compass-Card" Dial for Use in Taking Bearings from a Rotating Beacon.

the chronograph is observed at this instant of minimum signal intensity it is evident that the bearing of the transmitter from the receiver can be obtained by a simple calculation. To provide for the case in which the observer is due north or south of the beacon, when the *N* signal would probably be inaudible, another characteristic signal is emitted after a 90 deg. rotation to the corresponding *E* point. Bearings observed from this signal as a starting point are evidently subject to a correction of 90 deg. It is to be noted that since the radiation from the coil is symmetrical about its plane, a second minimum will be obtained

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after a rotation of 180 deg. from the first. With the beacon making one revolution per minute, therefore, a line bearing is obtainable in the above manner every half-minute. To fix the position of a receiving station it is necessary to obtain line bearings in this manner from two or more beacons.

Since the timing process mentioned above is but an intermediate step in taking a bearing, it is convenient to provide a stop-watch or chronograph used for the purpose, with a dial specially engraved in degrees and points of the compass as illustrated in Fig. 3. If the center second-hand of such a watch is started on the N signal from the beacon, the indication of the hand at the occurrence of the signal minimum will give the true bearing of the observer. Examples of other dials suitable for working from the E point, and on beacons with different times of rotation, have been described elsewhere.[44(a)]

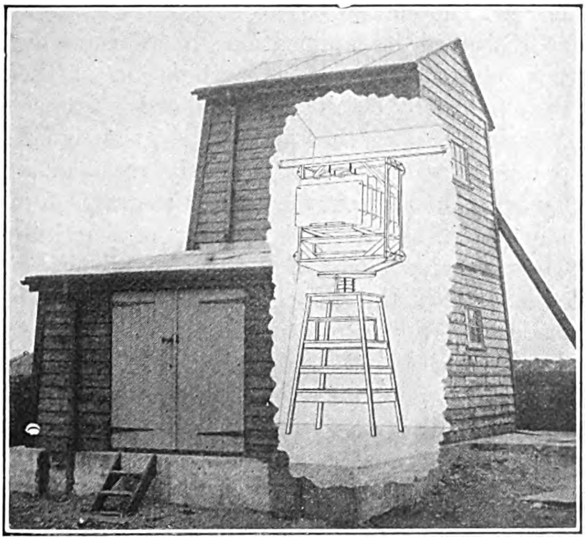

Fig. 4—View of Rotating Beacon inside Hut at Fort Monckton, Gosport.

The development and initial testing of this rotating beacon system by the Royal Air Force have naturally been confined to a study of its advantages over the direction-finding receiver for the navigation of aircraft.[30] During the past few years, however, the author has been privileged to investigate the performance of a rotating beacon station, especially set up for the purpose in England, and particularly to ascertain its advantages

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and reliability as an aid to marine navigation.[44] A sketch photograph of this beacon is shown in Fig. 4.

(c) *The Reversibility of Direction-Finding*

From the brief description of fundamental principles given in the two previous sections it will be gathered that wireless direction-finding can be carried out by rotating a closed loop at either the transmitting or receiving end of a communication link. The directional property is given by the rotating closed loop and the other station can employ any type of wireless transmitter or receiver connected to any type of aerial system. It will probably aid the reader in understanding the performance of each system and their relative merits for any given purpose, if attention is drawn here to the reversibility of the communication system between an open aerial and a closed loop.

The reference of the reciprocal theorem to radio communication has been enunciated by Lorentz, Pfrang, and Sommerfeld,[45] who discussed its application to transmission between open aerials, closed loops, or a combination of these. The theorem may be briefly stated in the following terms: if an antenna A_1 transmits to another antenna A_2, the signal intensity in A_2 is the same as that which would be received in A_1 if A_2 transmitted with the same power and at the same frequency as was previously used by A_1. This equality of signal intensity in the two cases is independent of the electromagnetic properties of the medium and of the shapes of the antennas. This means that transmissions from a rotating loop beacon being received on an open antenna should produce errors of the same type as those experienced when the antenna is employed for transmitting to a loop direction-finder. The utility of this theorem in studying the performance of the two systems will become evident in succeeding sections of the paper, and the author has referred elsewhere to the possibilities of investigating local errors in this manner.[46] On theoretical grounds, the reversibility of the Adcock direction-finding system as a means of avoiding night errors has also been established by the present author.[72] In the above cases a reservation must be made as to any irreversible effects which may be produced by the influence of the earth's magnetic field upon wireless waves travelling through ionized regions of the earth's atmosphere. The possibility of such a departure from the reciprocity relation was pointed out by Appleton in 1925.[47]

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TABLE I

Summary of Observations Taken at Newcastle from February 28, 1921 to March 4, 1922

(1)	(2)	(3)	(4)	(5)	(6)				(7)			
					Day Observations				Night Observations			
Transmitting Station	Type and wave-length	Time of trans-mission	Distance	True bearing	No.	Extent of variation	Error of mean	Max. varia-tion from mean	No.	Extent of varia-tion	Error of mean	Max. varia-tion from mean
	(km)		(miles)	(degs.)		(degs.)	(degs.)	(degs.)		(degs.)	(degs.)	(degs.)
Aberdeen	C.W. 3.3	I	154	352.0	100	10.5	−0.2	− 6.8	37	17.0	+0.1	−11.1
Aranjues	S. 3.5	I	1,034	184.6	65	1.7	+2.0	− 1.1	102	5.6	+1.6	± 2.8
Budapest	S. 3.0	I	1,026	111.6	45	3.5	−1.3	− 1.8	527	29.5	−1.1	+18.5
Chelmsford	C.W. 3.8	I	240	158.1	126	5.0	−0.8	+ 2.7	256	4.0	−0.7	+ 2.6
Cleethorpes	S. 3.0	17.00	118	147.2	—	—	—	—	87	4.2	−0.9	+ 2.2
Cleethorpes	C.W. 4.6	I	118	147.2	72	9.3	−0.8	− 5.2	134	12.0	−1.5	− 7.2
Clifden	S. 5.8	I	355	256.2	64	2.2	−1.4	+ 2.8	4	6.0	−3.4	+ 3.7
Clifden	C.W. 5.8	I	355	256.2	95	7.5	−0.3	+ 6.1	182	33.0	−0.1	−18.1
Coltano	S. 4.2	I	948	140.6	141	3.1	−1.9	+ 1.7	60	7.6	−0.7	+ 5.7
Horsea	C.W. 6.0	12.00	287	175.5	204	6.5	−0.9	+ 3.4	392	14.5	−0.5	+ 8.0
	4.5	20.00										
Karlsborg	S. 2.5	12.15	650	62.4	33	6.7	−1.4	+ 3.5	441	37.0	+1.7	+25.9
		20.00										
Karlsborg	S. 4.2	I	650	62.4	207	10.2	−1.2	− 6.4	128	23.0	+0.7	−12.1
Karlsborg	C.W. 4.5	I	650	62.4	27	2.3	−0.6	− 1.3	9	4.0	−2.3	+ 2.4
Moscow	S. 4.8	I	1,521	71.8	18	2.4	−0.4	− 1.3	1,838	34.0	+0.8	+23.4
Nantes	S. 2.5	08.00	538	180.4	2	0.3	0.0	− 0.2	8	7.6	+0.2	− 4.4
		21.00										
Nantes	S. 4.2	08.15	538	180.4	6	1.3	−2.9	+ 0.9	5	2.0	+0.4	+ 1.2
		21.30										
Nauen	S. 3.9	11.50	612	99.2	269	3.4	−2.5	− 2.0	—	—	—	—
	3.2											
Nauen	C.W. 4.7	I	612	99.2	50	13.5	−3.1	+ 6.9	27	13.5	−2.4	− 7.3
Nicolaieff	S. 4.0	I	1,546	96.3	160	5.8	−0.7	+ 4.4	106	12.8	+1.0	+ 7.2
Paris	S. 2.6	08.15	454	156.9	202	3.0	−2.2	− 1.7	—	—	—	
Paris	S. 2.6	09.20	454	156.9	1138	4.3	−2.2	− 2.9	—	—	—	
		10.40										
Paris	S. 2.6	11.30	454	156.9	605	3.1	−2.3	− 1.7	—	—	—	
Paris	S. 2.6	19.20	454	156.9	—	—	—		448	8.5	−2.1	+ 4.7
		21.10										
Paris	S. 3.2	11.00	454	156.9	537	3.0	−0.5	− 1.9	—	—	—	
		12.00										
Poldhu	S. 2.8	09.30	374	205.6	365	5.2	+0.3	+ 2.8	172	32.5	−1.0	−20.6
		21.30										
Poldhu	S. 2.8	I	374	205.6	14	1.5	+0.3	− 0.9	17	26.2	−4.5	−21.1
Sofia	S. 3.2	I	1,400	116.8	7	2.8	−2.2	± 1.4	166	11.5	+0.4	+ 5.8
Warsaw	S. 2.0	I	942	92.0	60	4.0	+0.8	+ 2.2	23	5.6	+0.2	+ 3.3

I = Irregular transmission.

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3. Analysis of the Performance of Wireless Direction-Finders

Attention has already been drawn in Section 1 to the carrying out of a large investigation in wireless direction-finding by the Radio Research Board in the British Isles during the years 1920-1926. In the course of this investigation, thirteen direction-finding stations were used in various parts of the work, which covered a range of wavelengths of from 300 to 20,000 meters. The precise distribution of these stations and the detailed results of the whole investigation will be found in a series of official reports to which reference has already been made.[16] It is considered to be useful here, however, to give a very brief resumé of these results with the conclusions drawn therefrom. Apart from the errors due to local conditions at each direction-finding station, which were studied separately and which could be corrected for when necessary, it was found that the apparent bearings of the various transmitting stations observed varied in a very erratic manner under certain conditions. In the work of tabulating and correlating the results it was found convenient to adopt an arbitrary division of the times of observation into day and night periods. The border lines between these periods were taken at one hour after sunrise at the westerly end of the path of transmission and one hour before sunset at the easterly end.

(a) *General Nature of Variations in Observed Bearings*

As illustrating the results generally obtained, a summary of the observations recorded at Newcastle on wavelengths between 2,500 and 6,000 meters during the years 1921–23 is reproduced in Tables I and II, while in Table III is reproduced a summary of the results obtained at Orford from 1922 to 1924 on wavelengths of 450 and 600 meters. With the mode of separation into day and night periods thus adopted it was found that the extreme error experienced in the day periods was usually about 4 deg. for any wavelength and distance of transmission, but during the winter months this limit was considerably exceeded on the higher wavelengths, apparently because of an extension of the night conditions until three or four hours after sunrise. As an example of this phenomenon Fig. 5 shows graphs of the daily bearings observed at Slough on the transmissions from the two stations at Leafield and Nantes. The observations were made on the U.R.S.I. signals sent by these stations at 1400 and 1415 G.M.T.,

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TABLE II

Summary of Observations Taken at Newcastle from March 6, 1922 to March 31, 1923

(1)	(2)	(3)	(4)	(5)	(6)				(7)			
					Day Observations				Night Observations			
Transmitting Station	Type and wave-length	Time of trans-mission	Distance	True bearing	No.	Extent of variation	Error of mean	Max. varia-tion from mean	No.	Extent of varia-tion	Error of mean	Max. varia-tion from mean
	(km)		(miles)	(degs.)		(degs.)	(degs.)	(degs.)		(degs.)	(degs.)	(degs.)
Aberdeen	C.W. 3.3	08.30–09.30 17.00	154	352.0	189	17.5	+0.7	−11.4	125	33.8	+0.8	+18.0
Aberdeen	C.W. 3.3	21.00	154	352.0	—	—	—	—	111	37.2	+0.5	−19.7
Air Ministry	C.W. 4.1	I	252	162.5	247	10.3	−1.2	+ 6.9	252	30.7	−0.8	+17.5
Berne	C.W. 3.4	I	676	140.8	530	5.1	−1.5	− 4.3	217	56.0	−0.6	+34.3
Budapest	S. 3.0	I	1,026	111.6	48	4.0	−2.0	+ 2.4	26	7.1	0.0	+ 4.0
Chelmsford	C.W. 3.8	I	240	158.1	76	4.4	−1.6	+ 2.9	70	2.5	−1.9	+ 1.3
Cleethorpes	C.W. 4.3	08.00	118	147.2	118	12.0	−1.5	− 6.2	—	—	—	—
Cleethorpes	C.W. 4.3	I	118	147.2	116	8.5	−2.3	− 4.9	187	21.0	−0.3	−11.9
Clifden	S. 5.8	09.50 21.50	355	256.2	85	1.5	−1.4	− 0.8	66	12.5	−2.5	− 7.7
Clifden	C.W. 5.8	I	355	256.2	316	2.3	−0.5	+ 1.4	594	26.0	−0.4	+14.2
Coltano	S. 4.2	I	948	140.6	169	3.3	−1.9	− 2.2	244	10.8	−1.6	± 5.4
Horsea	C.W. 5.0	I	287	175.5	193	13.6	−0.8	+ 8.8	98	15.6	−1.3	− 9.3
Karlsborg	S. 2.5	12.15 20.00	650	62.4	171	6.6	+0.1	+ 3.5	485	57.5	+5.1	−31.5
Karlsborg	S. 4.2	I	650	62.4	125	6.2	−0.5	− 4.3	53	26.3	+0.1	+15.0
Karlsborg	C.W. 3.9	I	650	62.4	503	10.5	+2.2	− 7.1	237	55.0	+3.8	+38.8
Konigswusterhausen	C.W. 5.2	I	658	100.0	501	8.3	−2.6	− 4.2	260	32.4	−3.3	−20.7
Moscow	S. 4.8	I	1,521	71.8	421	5.5	−1.6	+ 2.8	945	39.0	−0.8	+26.5
Nantes	S. 2.5	08.00 21.00	538	180.4	21	0.6	−0.8	+ 0.4	94	11.9	−0.2	− 6.6
Nantes	S. 4.2	08.10 21.30	538	180.4	84	4.5	−0.2	− 3.0	93	15.7	−1.0	+10.6
Nantes	S. 2.5	12.15	538	180.4	60	5.2	−2.4	− 3.0	—	—	—	—
Nauen	S. 3.2	11.50	612	99.2	250	4.0	−2.8	− 2.4	—	—	—	—
Nauen	C.W. 4.7	I	612	99.2	1421	87.0	−2.5	+60.8	1471	78.2	−0.8	−40.4
Ongar (GLA)	C.W. 2.9	I	238	160.2	904	11.0	−0.1	− 6.1	979	59.5	−0.2	+30.0
Ongar (GLB)	C.W. 3.8	I	238	160.2	1239	7.5	−0.8	− 9.1	859	34.4	−1.2	−17.4
Ongar (GLO)	C.W. 4.4	I	238	160.2	252	9.5	−1.4	+ 5.7	321	23.9	−1.7	−13.0
Paris	S 2.6	08.15	454	156.9	277	2.4	−2.8	+ 1.4	—	—	—	—
Paris	S. 2.6	09.20 19.20	454	156.9	1486	5.5	−3.1	− 3.8	432	5.8	−2.7	− 4.0
Paris	S. 2.6	11.30 21.55	454	156.9	516	2.4	−2.7	+ 1.3	162	13.2	−4.0	− 7.[illegible]
Paris	S. 3.2	12.00	454	156.9	114	1.5	−1.1	− 0.8	—	—	—	—
Poldhu	S. 2.8	09.50 21.50	374	205.6	42	2.9	+0.4	− 1.5	37	30.0	−1.9	−21.7
Sofia	S. 3.2	I	1,400	116.8	—	—	—	—	163	48.0	−1.8	−31.5
Whitehall	C.W. 3.3	09.00 17.00	247	164.9	43	6.9	+0.9	+ 4.3	37	7.1	+1.8	+ 3.8
Whitehall	C.W. 3.3	21.00	247	164.9	—	—	—	—	30	12.0	−0.3	− 6.1

I = Irregular transmission.

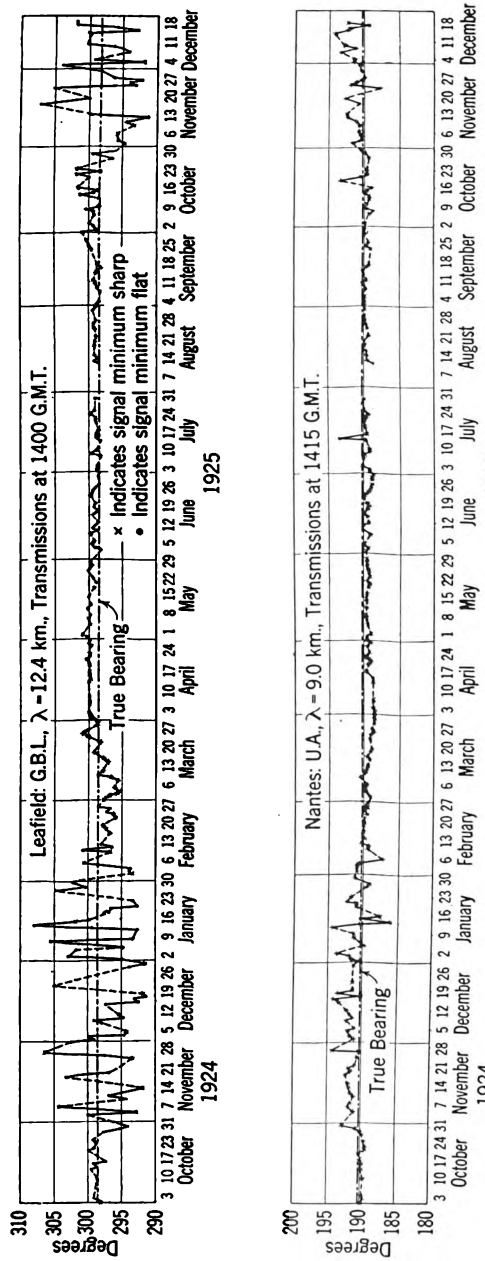

Fig. 5—Graphs Showing the Daily Observations of Bearings Taken at Slough on the U.R.S.I. Transmissions from Leafield and Nantes. October, 1924 to December, 1925.

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respectively, and thus always took place in daylight. The results are plotted as the weekly extreme bearings over a period of fifteen months from October, 1924 to December, 1925. It is evident from this diagram that during the summer months the observed bearings are steady and fairly accurate, while during

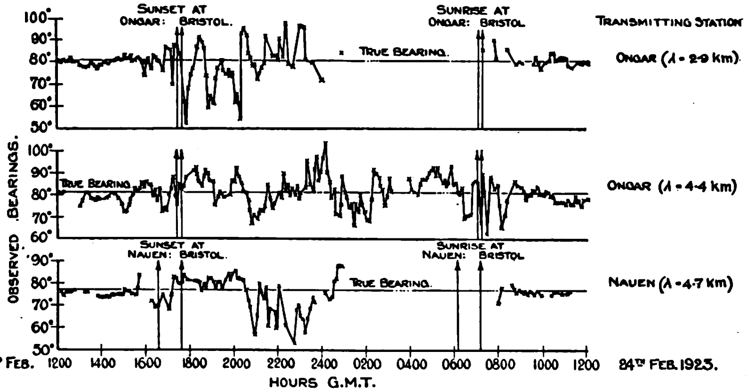

Fig. 6—Graphs of Observed Bearings of Ongar (undamped waves, λ = 2.9 and 4.4 km) and Nauen (undamped waves, λ = 4.7 km) taken at Bristol over a 24-hour period.

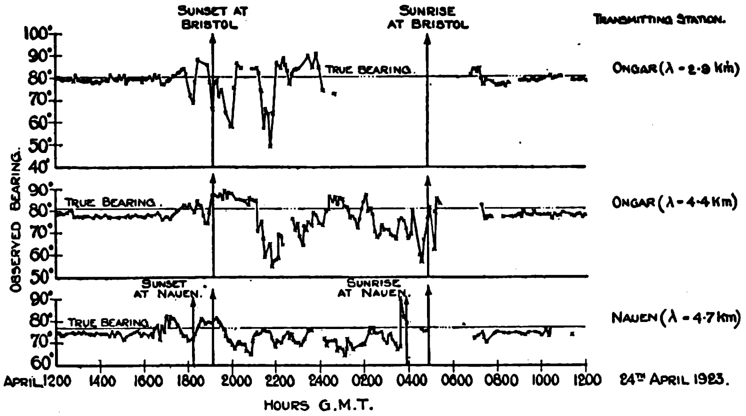

Fig. 7—Graphs of Observed Bearings of Ongar (undamped waves, λ = 2.9 and 4.4 km) and Nauen (undamped waves, λ = 4.7 km) taken at Bristol over a 24-hour period.

the winter months the daily errors assume appreciable proportions, particularly in the case of Leafield.

The general nature of the variations can be understood from the graphs given in Figs. 6 and 7, which show the apparent bearings of some fixed transmitting stations as observed every

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TABLE III
Summary of Observations Taken at Orford from November 13, 1922 to March 29, 1924

(1)	(2)	(3)	(4)	(5)				(6)			
				Day Observations				Night Observations			
Transmitting Station	Type and wavelength	Distance	True bearing	No.	Extent of variation	Error of mean	Max. variation from mean	No.	Extent of variation	Error of mean	Max. variation from mean
	meters	(miles)	(degs.)		(degs.)	(degs.)	(degs.)		(degs.)	(degs.)	(degs.)
Amsterdam	S. 450	143	81.1	46	3.0	+0.6	−1.7	9	1.0	+1.2	+ 0.7
Antwerp	S. 600	136	114.9	152	3.0	+0.5	+1.6	94	30.0	+0.9	−15.8
Berwick	S. 450	285	332.0	161	2.5	+0.2	+2.3	106	20.0	+0.6	+12.4
Borkum	S. 450	237	62.3	65	3.0	+0.3	−1.6	127	14.0	+0.1	− 7.4
Boulogne	S. 450	95	178.2	519	3.5	−0.3	+2.1	341	15.0	0.0	+11.8
Cherbourg	S. 450	220	219.9	344	5.5	−0.2	+2.8	294	90.0	+1.0	+59.1
Cullercoats	S. 600	236	330.0	142	4.0	−0.3	−1.7	110	45.0	+0.5	+24.5
Dieppe	S. 450	151	188.0	206	6.0	+0.4	−3.4	—	—	—	—
Dunkirk	S. 450	83	154.1	294	3.0	+0.4	±1.5	193	32.0	+0.9	+20.0
Dunkirk	S. 600	83	154.1	90	2.0	+0.4	±1.0	85	15.0	+0.3	− 9.4
Flamborough	S. 450	155	334.8	331	2.5	−0.4	−1.4	258	45.0	−1.5	−28.3
Flushing	S. 450	98	116.0	936	4.0	+0.7	+2.3	173	30.0	+0.6	−16.6
Grimsby	S. 600	123	328.0	149	4.0	−0.5	+2.5	212	20.0	+0.8	+11.2
Havre	S. 450	188	200.0	977	5.5	−0.6	+3.1	563	75.0	+1.5	+58.5
Havre	S. 600	188	200.0	25	3.0	−0.6	−1.6	—	—	—	—
Helder	S. 450	148	65.0	544	4.0	−0.3	−2.2	108	11.0	+0.2	+ 9.8
Lizard	S. 450	327	246.2	—	—	—	—	194	25.0	+2.7	−13.9
Mok	S. 450	149	64.0	155	3.0	−1.3	−1.7	—	—	—	—
Niton	S. 600	161	230.5	81	5.5	−4.3	−3.2	40	20.0	−4.6	−15.9
North Foreland	S. 600	49.5	185.8	380	2.5	−1.2	+2.4	404	15.0	−1.2	+10.4
Ostend	S. 600	84	135.2	312	4.0	+0.1	+1.7	232	24.0	+0.1	+13.7
Parkeston Quay	S. 450	16.5	230.5	401	3.0	−2.9	−1.6	350	4.0	−3.2	− 2.3
Parkeston Quay	S. 600	16.5	230.5	10	0.5	−2.8	+0.3	10	1.0	−2.9	− 0.6
Portland	S. 450	204	239.4	20	7.0	−0.5	−3.9	47	35.0	+7.6	−22.0
Rochefort	S. 600	440	195.8	—	—	—	—	106	20.0	+2.8	+11.4
Scheveningen	S. 600	117	89.2	110	6.5	−0.5	+3.8	52	25.0	−1.1	−13.1
Scheveningen	S. 1,800	117	89.2	139	5.5	−0.4	−3.8	68	35.0	−0.3	−18.9
Teddington	S. 450	93	240.6	712	4.5	−2.6	+2.5	551	125.0	+1.7	+82.7
Teddington	S. 500	93	240.6	72	2.5	−3.2	−1.4	135	13.0	−3.1	+12.5
Teddington	S. 750	93	240.6	129	2.5	−3.1	−2.0	498	25.0	−2.7	+12.1
Teddington	I.C.W. 750	93	240.6	134	2.5	−3.0	+1.9	264	17.0	−2.7	+ 9.1
Teddington	C.W. 750	93	240.6	147	2.5	−3.1	+1.5	444	20.0	−2.4	+11.8
Teddington	S. 1,000	93	240.6	34	0.2	−2.6	+0.2	101	0.5	−3.0	+ 0.4
Teddington	I.C.W.1,000	93	240.6	34	1.0	−2.6	±0.5	76	2.2	−3.1	+ 2.2
Teddington	C.W. 1,000	93	240.6	134	2.0	−2.5	±1.4	221	10.0	−2.1	+ 6.5
Teddington	S. 1,800	93	240.6	92	2.5	−1.8	−1.3	120	6.0	−2.1	− 3.5
Teddington	C.W. 1,800	93	240.6	196	3.5	−1.7	+2.1	328	30.0	−1.4	+15.8
Teddington	C.W. 2,600	93	240.6	65	2.5	−1.4	−1.7	123	35.0	−2.1	+21.5
Treguier St. Gonery	S. 450	307	224.9	15	3.5	−2.6	−2.3	12	13.0	+3.9	− 6.8
Ushant	S. 450	386	231.9	19	5.0	+0.2	+2.9	548	60.0	+2.0	+36.1
Wilhelmshaven	S. 450	293	67.8	304	4.0	−0.6	−2.2	323	35.0	−1.9	+19.1

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few minutes over periods of 24 hours at one observing station. It is to be noticed that the day bearings are much steadier in the summer than in the winter months, but that in either case the approach of sunset is accompanied by an increase in the magnitude and frequency of the variable errors which continues throughout the night until sunrise. During the bulk of the investigations carried out on wavelengths between 450 and 12,000 meters these variable errors in bearing ranged up to, but very rarely exceeded, 90 deg. as illustrated in the above tables.

In some of the later work the observations have been extended into the broadcasting band of wavelengths. A summary of the results obtained is given in Table IV, and this shows that in the

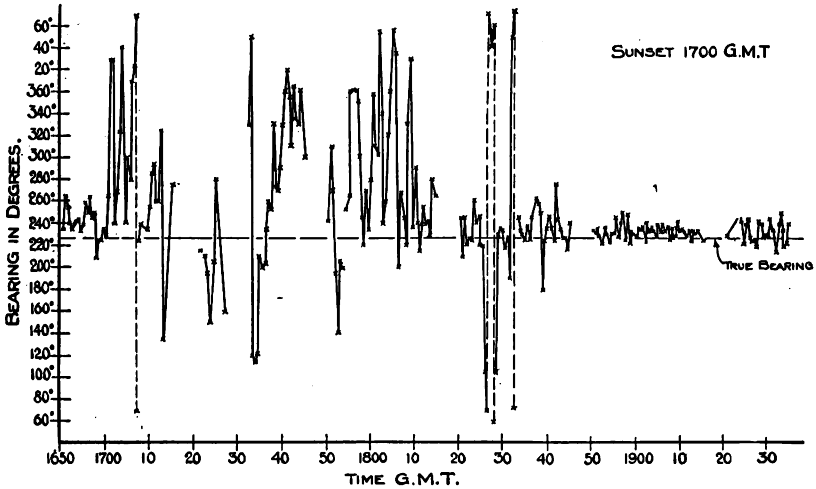

Fig. 8—Graphs of Observed Bearings of Bournemouth (undamped waves, $\lambda = 386$ m) taken at Ditton Park, Slough. October 20, 1926.

case of the observations at Slough taken on the transmissions from Bournemouth, the maximum error is given as ± 180 deg. This indicates that a single minimum has been followed round the direction-finder scale through this angle and indeed on several occasions a rotation of the apparent bearing through more than 360 deg. has been observed. Instances of this effect are illustrated in Fig. 8. A somewhat similar phenomenon has been previously recorded by Wright and Smith,[48] but on the much longer wavelength of 6,000 meters.

Another mode of illustrating the variable errors due to night conditions is that adopted in Fig. 9. This diagram refers to

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TABLE IV

Summary of Observations Taken at Slough from May 28 to December 17, 1925

(1)	(2)	(3)	(4)	(5)				(6)			
				Day Observations				Night Observations			
Transmitting Station	Wave-length	Distance	True bearing	No.	Extent of variation	Error of mean	Max. variation from mean	No.	Extent of variation	Error of mean	Max. variation from mean
	(meters)	(miles)	(degs).		(degs.)	(degs.)	(degs.)		(degs.)	(degs.)	(degs.)
Birmingham	479	88	319.1	201	3.0	+2.0	+1.9	1378	105.0	+3.4	−67.5
Bournemouth	386	77	227.6	76	6.2	+0.4	+4.2	1346	>360.	+5.3	±180.
London	365	18	83.5	—	—	—	—	803	7.5	+1.3	− 4.8
Newcastle	404	245	350.2	—	—	—	—	537	75.0	−1.4	+51.4

(The transmissions were all on telephony of which the carrier wave was treated as c.w. for reception purposes.)

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TABLE V

Showing Percentage Variations of Bearings from the Annual Mean at Newcastle, 1922–23.

(1)	(2)		(3)	(4)		(5)		(6)		(7)		(8)	
Transmitting Station	Type and wave-length		Time of trans-mission	Total number of observations		Percentage Variations from Annual Mean							
						Over 2 deg.		Over 5 deg.		Over 10 deg.		Over 20 deg.	
		(km)		Day	Night	Day	Night	Day	Night	Day	Night	Day	Night
Aberdeen	C.W.	3.3	08.30 17.00	189	125	37.0	63.1	6.3	28.0	0.5	13.6	0.0	0.0
Aberdeen	C.W.	3.3	21.00	—	111	—	82.0	—	49.5	—	10.1	—	0.0
Air Ministry	C.W.	4.1	I	247	252	17.0	33.3	0.8	8.3	0.0	1.6	0.0	0.0
Berne	C.W.	3.4	I	530	217	3.2	61.3	0.4	32.7	0.0	10.6	0.0	1.8
Budapest	S.	3.0	I	48	26	2.1	15.4	0.0	0.0	0.0	0.0	0.0	0.0
Chelmsford	C.W.	3.8	I	76	70	9.2	0.0	0.0	0.0	0.0	0.0	0.0	0.0
Cleethorpes	C.W.	4.3	08.00	118	—	30.5	—	2.5	—	0.0	—	0.0	—
Cleethorpes	C.W.	4.3	I	116	187	12.9	61.0	0.0	13.9	0.0	0.5	0.0	0.0
Clifden	S.	5.8	09.50 21.50	85	66	0.0	28.8	0.0	9.1	0.0	0.0	0.0	0.0
Clifden	C.W.	5.8	I	316	594	0.0	6.9	0.0	3.4	0.0	1.2	0.0	0.0
Coltano	S.	4.2	I	169	244	1.0	16.4	0.0	1.0	0.0	0.0	0.0	0.0
Horsea	C.W.	5.0	I	193	98	23.8	39.8	2.6	10.2	0.0	0.0	0.0	0.0
Karlsborg	S.	2.5	12.15 20.00	171	485	20.5	83.5	0.0	42.5	0.0	20.2	0.0	1.9
Karlsborg	S.	4.2	I	125	53	6.4	50.9	0.0	18.9	0.0	7.5	0.0	0.0
Karlsborg	C.W.	3.9	I	503	237	11.3	40.0	0.6	16.9	0.0	8.5	0.0	2.1
Konigswusterhausen	C.W.	5.2	I	501	260	14.2	49.6	0.0	18.1	0.0	4.2	0.0	0.4
Moscow	S.	4.8	I	421	945	0.5	21.9	0.0	6.6	0.0	1.7	0.0	0.2
Nantes	S.	2.5	08.00 21.00	21	94	0.0	26.6	0.0	5.3	0.0	0.0	0.0	0.0
Nantes	S.	4.2	08.10 21.30	84	93	2.4	35.5	0.0	8.6	0.0	1.1	0.0	0.0
Nantes	S.	2.5	12.15	60	—	6.7	—	0.0	—	0.0	—	0.0	—
Nauen	S.	3.2	11.50	250	—	2.0	—	0.0	—	0.0	—	0.0	—
Nauen	C.W.	4.7	I	1421	1471	17.7	52.7	6.8	23.9	2.2	5.9	0.3	0.6
Ongar	C.W.	2.9	I	904	979	19.3	50.2	0.1	24.9	0.0	5.9	0.0	0.6
Ongar	C.W.	3.8	I	1239	859	2.1	34.7	0.1	14.8	0.0	2.7	0.0	0.0
Ongar	C.W.	4.4	I	252	321	4.7	24.0	0.4	10.6	0.0	1.6	0.0	0.0
Paris	S.	2.6	08.15	277	—	0.0	0.0	0.0	0.0	0.0	0.0	0.0	0.0
Paris	S.	2.6	09.20 19.20	1486	432	3.0	1.4	0.0	0.0	0.0	0.0	0.0	0.0
Paris	S.	2.6	11.30 21.55	516	162	0.0	23.4	0.0	2.5	0.0	0.0	0.0	0.0
Paris	S.	3.2	12.00	114	—	0.0	—	0.0	—	0.0	—	0.0	—
Poldhu	S.	2.8	09.50 21.50	42	37	0.0	65.0	0.0	27.0	0.0	8.1	0.0	5.4
Sofia	S.	3.2	I	—	163	—	45.4	—	16.5	—	3.7	—	0.6
Whitehall	C.W.	3.3	09.00 17.00	43	37	30.2	46.0	0.0	0.0	0.0	0.0	0.0	—
Whitehall	C.W.	3.3	21.00	—	30	—	40.0	—	10.0	—	0.0	—	0.0

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records by several observers of the apparent bearings of Karlsborg during its daily transmissions from 2000 to 2030 on a wavelength of 2,500 meters. The observations were summarized in weekly batches and the graphs show the extremes of the bearings for individual weeks over a period of two years. It is very noticeable that the bearings show little variation so long as they are taken before sunset at the receivers, even though

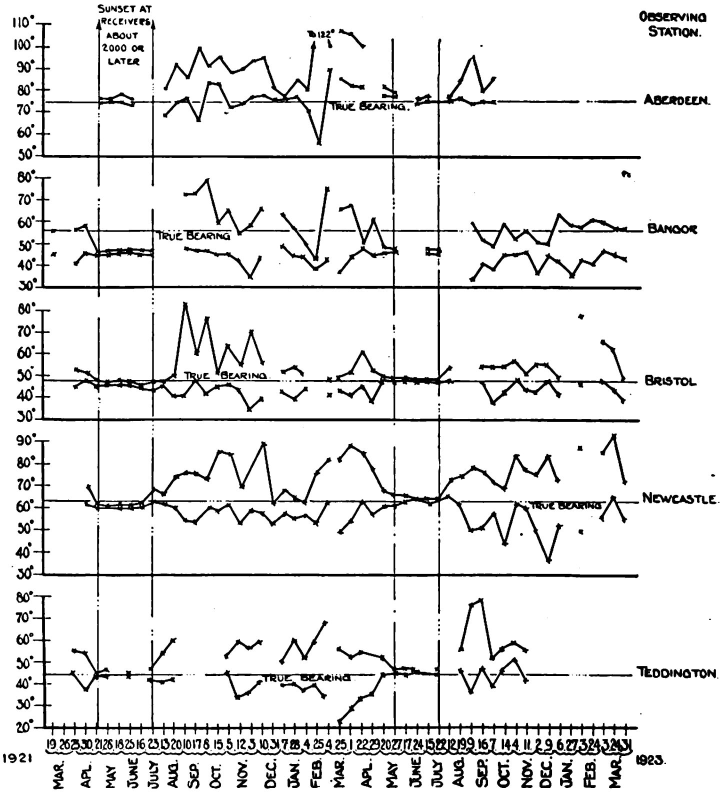

Fig. 9—Graphs of Weekly Extreme Bearings of Karlsborg (damped waves, $\lambda = 2.5$ km) observed on the transmissions at 2000 G.M.T. during the two years, March, 1921 to March, 1923.

darkness has prevailed over a portion of the path of transmission for some time.

In spite of the large variations in apparent bearings described and illustrated above, it is noteworthy that during the night periods the observed bearings show no signs of a definite system-

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atic error, and so the variations are on the whole equally distributed about the mean value. Another feature to be observed in connection with the magnitude of the night variations is the comparative rarity of the larger errors. A summary of the results obtained at Newcastle during the last year of operation is given in Table V, arranged in such a manner as to show the proportion of the observed bearings which differ by various fixed amounts from the mean day bearing. From this mode of expressing the results it will be seen that it is not often that more than 10 per cent of the night results give an error exceeding 10 deg. This fact, combined with the absence of a systematic error at night, is of great importance in the practical application of wireless direction-finding.

It thus appears that the chief effects to be observed in radio direction-finding are generally as follows. In the summer time in those latitudes in which the United Kingdom is situated the observed bearings in daylight are fairly constant and, with the exception of the permanent errors already mentioned, they approximate quite accurately to the true geographical bearing of the transmitting station. Under certain conditions of transmission to be referred to below this accuracy is maintained at all times and seasons. When these conditions do not prevail, however, the observed bearings show signs of variable errors as the setting of the sun approaches the path of transmission. These variations then prevail throughout the night, but decrease to a negligible extent soon after sunrise. As the seasonal conditions are changed gradually from summer to winter the variations in the daytime increase somewhat; and the night conditions begin an appreciable time before sunset and continue until some time after sunrise. In the midwinter conditions represented by January, the change from day to night effects is very much more gradual and the difference is only distinguishable by the difference in magnitudes of the variations. At this time of year it appears that on the longer wavelengths the variations may have an amplitude of the order of 10 deg. or more during the daylight hours, and this increases rapidly at about one hour before sunset to the night value of 50 deg. or 70 deg., and so continuing in the most erratic manner until one hour or so after sunrise.

It has now been satisfactorily established that these variations in apparent bearing are due to the arrival at the receiver

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of waves from the upper atmosphere polarized with their electric force in a horizontal plane. This explanation was first put forward by T. L. Eckersley[49] in 1921 and supported by experimental results. Of recent years considerable direct confirmation of the theory has been provided,[50,51,52] and the study of the propagation of electromagnetic waves through the ionized regions of the atmosphere now forms the subject of large numbers of papers published in all parts of the world.

(*b*) *Effect of Various Factors upon Variations in Bearings*

From time to time throughout the investigation the results have been carefully considered to ascertain if there is any definite relation between the wavelength of transmission and the nature, frequency, and magnitude of the variations in apparent bearings experienced. Over the band of wavelengths from 300 to 20,000 meters no marked difference has been observed in the extreme variations recorded, after due allowance has been made for the distance and the geographical conditions over the path of transmission. The outstanding exception to this statement is the case of Bournemouth observed at Slough, as mentioned above, in which exceedingly violent variations have been recorded. It is probable that this occurrence is due to a coincidence of wave length and distance, and the relative magnitudes of the downcoming and direct waves. The conclusion that, within wide limits, the wavelength has no effect upon the amplitude of the variations observed over a period is to be distinguished from the fact that when the wavelength of transmission is changed instantaneously the error in bearing due to night effect also changes appreciably. Excellent examples of this occurrence have been obtained in observing on the transmissions from arc stations, such as Leafield, which employ for signalling purposes a marking and spacing wave differing in length by about 1 per cent. The errors in bearing taken in rapid succession on the marking and spacing waves differed at times by 20 deg. to 30 deg.

The experience of the author obtained in Great Britain that the wavelength itself does not have a very marked effect upon the magnitude of the variations would appear to differ from that of Austin,[53] who states that in the United States of America the variations on the shorter wavelengths are considerably less than on the longer wavelengths. This conclusion does not, however, appear to be consistent with the results obtained by

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TABLE VI

Summary of Observations Taken at Teddington during 24-hour Test from January to June, 1926

(1)		(2)	(3)	(4)	(5)	(6)	(7)	(8)	(9)			
Transmitting Station		Wave-length	Distance	True bearing	No. of obser-vations	Extent of varia-tion	Error of mean bearing	Max. varia-tion from mean	Percentage of Bearings differing from Mean by more than:			
									2 deg.	5 deg.	10 deg.	20 deg.
		(meters)	(miles)	(degrees)		(degs.)	(degrees)	(degs.)				
Bordeaux	LY	18 940	452	181.8	316	33.0	−0.2	−19.1	63.3	32.6	5.4	0.0
Carnarvon	MUU	14 200	200	307.3	139	7.7	+1.3	− 4.6	36.9	0.0	0.0	0.0
Karlsborg	SAQ	18 050	760	44.4	219	19.8	+2.9	−11.6	55.7	11.9	0.9	0.0
Long Island	WQK	16 460	3,403	287.9	547	7.4	+2.3	+ 3.4	6.4	0.0	0.0	0.0
Rugby	GBR	18 900	75	330.4	147	34.5	−1.3	+27.9	85.8	47.7	11.6	4.8
St. Assise	UFT	14 290	237	146.3	416	34.0	−3.9	−22.4	83.7	35.4	9.4	0.3
Tuckerton	WCI	16 800	3,521	287.3	192	3.7	+2.4	+ 2.0	0.0	0.0	0.0	0.0

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Merritt,[54] Bidwell,[55] and Reich[56] at Cornell University, which show variations in bearings exceeding 100 deg. on wavelengths of 380 to 500 meters. The difficulty of drawing very definite conclusions from observations made on available transmissions from commercial stations within range is illustrated by Table VI, which is intended to show the decrease in magnitude of the bearing variations when the range of transmission is over 3,000 miles. The results given in this table were obtained at Teddington, England, during a series of continuous tests each lasting 24 hours, in which observations were made of the apparent bearings of various European and American stations transmitting on wavelengths between 14,000 and 19,000 meters. These experiments have shown that over similar distances the variations experienced on the wavelengths 2,000 to 9,000 meters are of the same order as those on the higher wavelengths, and also that the actual amplitudes of the variations decrease considerably for a range of transmission exceeding 3,000 miles. In fact, on the American transmitting stations the variations in

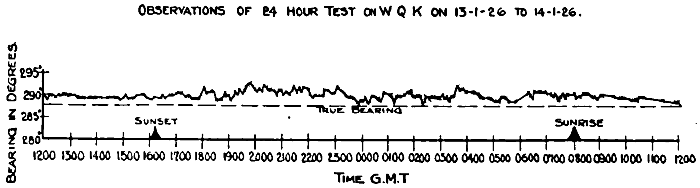

Fig. 10—Graph of Observed Bearings of Long Island (WQK, $\lambda = 16.5$ km) taken at Teddington during a 24-hour period, January 13 to 14, 1926.

bearings show no marked distinction between day and night periods. As illustrated in Table VI, observations taken over several 24-hour periods show a maximum error in bearing of about 3 deg. for the American stations, while on nearer European stations the night error may range up to nearly 30 deg. From Table VI it is seen that of the two American stations observed one gave 94 per cent and the other 100 per cent of bearings correct within 2 deg. during the whole series of tests; whereas of the European stations the best example, Carnarvon, gave only 63 per cent of bearings correct to within this limit. The contrast of the observations on the stations at different distances is illustrated by Figs. 10 and 11 which are typical of the results obtained on transmissions from the American and European stations. The error of the daylight readings is in each case attributable to local conditions at the direction-finder.

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While discussing the effect of distance on the errors in observed bearings it may be mentioned that Pickard[57] published in 1922 some results, of which a very noticeable feature was that the apparent bearings of European stations observed at Maine, U. S. A., showed much smaller variations than the bearings of some American stations. This difference was attributed to the fact that the transmission from the former was mostly over sea, whereas in the latter case it was entirely over land. It can be shown from theoretical considerations, however, that this is probably not the correct explanation.[16(d)] Mesny[58] has also found that the bearings on some American stations observed in France show only a very small variation as compared with bearings taken on the nearer European stations. Experiments carried out in Shanghai by Gherzi[59] during 1923–24 have also shown that while large night errors are observed for distances of transmission

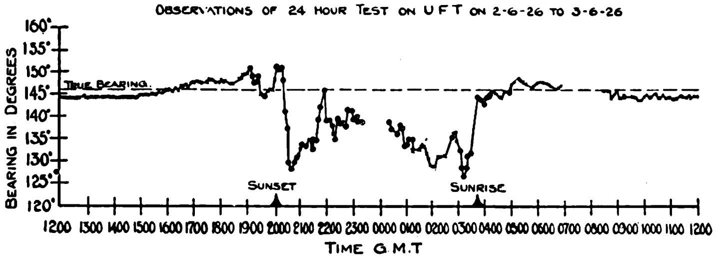

Fig. 11—Graphs of Observed Bearings of Ste. Assise (UFT, $\lambda = 14.3$ km) taken at Teddington during a 24-hour period, June 2 to 3, 1926.

of less than 5,000 km (3,000 miles) these errors have an extreme value of about 6 deg. when the range of transmission is from 5,000 to 12,000 km (3,000 to 7,500 miles).

The effect of the incidence of sunlight upon the upper regions of the earth's atmosphere can also be demonstrated during a solar eclipse. Fig. 12 shows the results of observations made at Slough on the transmissions from the Manchester broadcasting station during the solar eclipse of June 29, 1927. Although the normal night variations had ceased at about one hour after sunrise, it is seen that the variations were temporarily restored during the period of obscuration of the sun. A discussion of these and other wireless observations made during the eclipse will be found in the official published report.[60]

In concluding this section it may be repeated that it is only intended to be a general summary of the results obtained from

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the co-ordination of a large quantity of data. Anyone who is interested in obtaining more exact information on any portion of the work may be referred to the official reports containing the tabulated results in detail and published by the Radio Research Board. As far as the actual nature and magnitude of the effects observed are concerned, the author believes he is correct in stating that his results obtained in the British Isles are in complete agreement with the observations made in other countries, such as by Mesny and others in France;[58,61,62] by Stoye in Germany;[63] Austin,[64] Bidwell,[54,55] and Pickard[57] in America; and by Gherzi[59] in China.

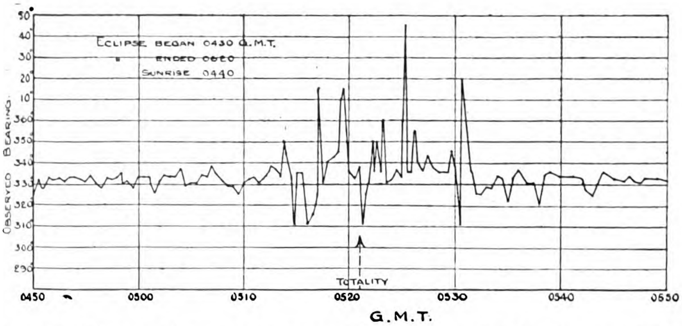

Fig. 12—Observations of Apparent Bearings of Manchester (2ZY) at Slough during the Solar Eclipse, June 29, 1927.

4. Analysis of the Performance of a Rotating Beacon

If the application of the principle of reversibility in direction-finding as stated in Section 2 be accepted, then it is evident that the performance of a rotating loop beacon transmitter can be largely predicted from the results and experience obtained with receiving loop direction-finders. Thus a rotating loop beacon when erected on the same site as a direction-finder will give bearing observations at a distant receiver, which will be subject to the same type of local error and night variations, for example, as the bearings observed on the direction-finder when the distant receiving aerial is used for transmission. Such deductions as these have been confirmed by the author during the past two or three years in the investigation of the performance of a rotating beacon erected at Fort Monckton, near Gosport, England.[44] Bearings observed on this beacon in certain directions were found to be subject to a small permanent error, which was of an approximately quadrantal nature, and which decreased in

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magnitude with increase of the working wavelength.[46] This error was of the same order as that experienced with a wireless direction-finder set up in proximity to the beacon, and was found to be most probably due to some underground power cables. In the choice of sites for future beacons it is evident that a portable direction-finder can be usefully employed in ascertaining the suitability of the site and its liability to produce local errors.

In order to ascertain the reliability of this type of rotating beacon as an aid to marine navigation, a number of tests were carried out on ships crossing the English Channel between Southampton and Havre, and Southampton and Jersey. Using the ship's ordinary wireless receiver observations of the bearing of the beacon were made at intervals during each trip and compared with the bearing as given by the Captain of the ship.

A typical log of one of these tests is given in Table VII, while the chart in Fig. 13 shows a comparison between the ship's "dead reckoning" and the wireless course in three other tests. As a result of tests conducted on these lines it was found that in the majority of cases the estimated and observed bearings agreed within from 2 deg. to 4 deg., although at times this difference ranged up to 12 deg. Signs of night effects in the shape of indistinct signal minima and wandering bearings were observed at distances exceeding 50 miles, but these were not always coincident with the above differences which sometimes occurred in daylight. In many cases at night and during misty weather, when visibility was very poor, the ship had to be navigated by dead reckoning so that the estimated bearing of the beacon from the ship may be subject to some suspicion.

In order to study these night variations in a more satisfactory manner, a number of tests were arranged in which observations were made continuously over a period of twelve hours or more from various fixed positions. The places of observation were selected so as to provide a variety of ranges and also to show the difference between transmission over sea and over land. Some of the observations were carried out in ships moored in dock or alongside a light vessel, while other observations with transmission entirely overland were carried out at Slough and Teddington.

An analysis of the readings obtained in this manner in daylight shows that for ranges up to 56 miles the maximum departure of the observed bearings from their mean or correct

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value was 3 deg., while up to the maximum range of 119 miles the greatest deviation was 6 deg. In most cases over 90 per cent of the observations were correct within 2 deg. When the observations were carried out at night, the above accuracies were maintained for distances overland of 14 miles and oversea of 23 miles. At distances of 92 miles entirely oversea the night errors experienced ranged up to a maximum value of 18 deg., but it is to be noted that in these cases over 84 per cent of the bearings were cor-

TABLE VII

Summary of Observations Obtained on a Return Voyage from Southampton to Havre, September 22 to 24, 1926.

Continuous Wave Transmission, $\lambda = 525$ m

Date	G.M.T.	From information supplied by Captain of ship				Wireless bearing from beacon	Remarks
		Position and how obtained		From Fort Monckton			
				Distance	Bearing		
				miles	degs.	degrees	
22.9.26	2240	Southampton.	A.	14.3	127	126	Mean of several readings.
23.9.26	0102	Beyond Nab.	V.B.	17.5	324	323	
	0132		D.R.	26.0	328	329	
	0202		D.R.	35.	330	331	
	0232		D.R.	43.	331	331	
	0302		D.R.	51.	332	332	
	0332		D.R.	60.	333	334	
	0532		D.R.	85.	335	335	
	0602		D.R.	94.	334	335 }	Night effect observed
	0632		D.R.	96.	330	331 }	
	0722	Havre.	A.	105.	328	330	
	1702	Havre	A.	105.	328	329	Mean of several readings.
24.9.26	0012		D.R.	95.	330	331	
	0032		D.R.	88.	330	329 }	Night effect observed.
	0102		D.R.	79.	329	327 }	
	0132		D.R.	69.	329	325	Night effect observed.
	0202		D.R.	60.	329	325	
	0232		D.R.	52.	329	325	
	0402		D.R.	24.	329	326	
	0432		V.B.	15.	325	326	
	0442	2 miles from Nab.	V.B.	11.5	324	323	

Notes.— (1) A = moored in dock, V.B. = visual bearings, D.R. = dead reckoning. (2) Night effect observed at 60 miles and over. (3) Maximum difference between estimated and observed bearings = 4 deg. at night. (4) Sea moderate throughout.

rect to within 5 deg. When the oversea distance exceeded 100 miles the errors became more serious. For overland transmission the errors were much larger, and at a distance of 55 miles errors up to 32 deg. were experienced, while only 66 per cent of the observed bearings were correct to within 5 deg. In spite of such large errors, however, it was found that the mean bearing of the beacon from any position was practically the same by day or night. This implies that the systematic error of the night variations is very small, so that the effect of errors of individual ob-

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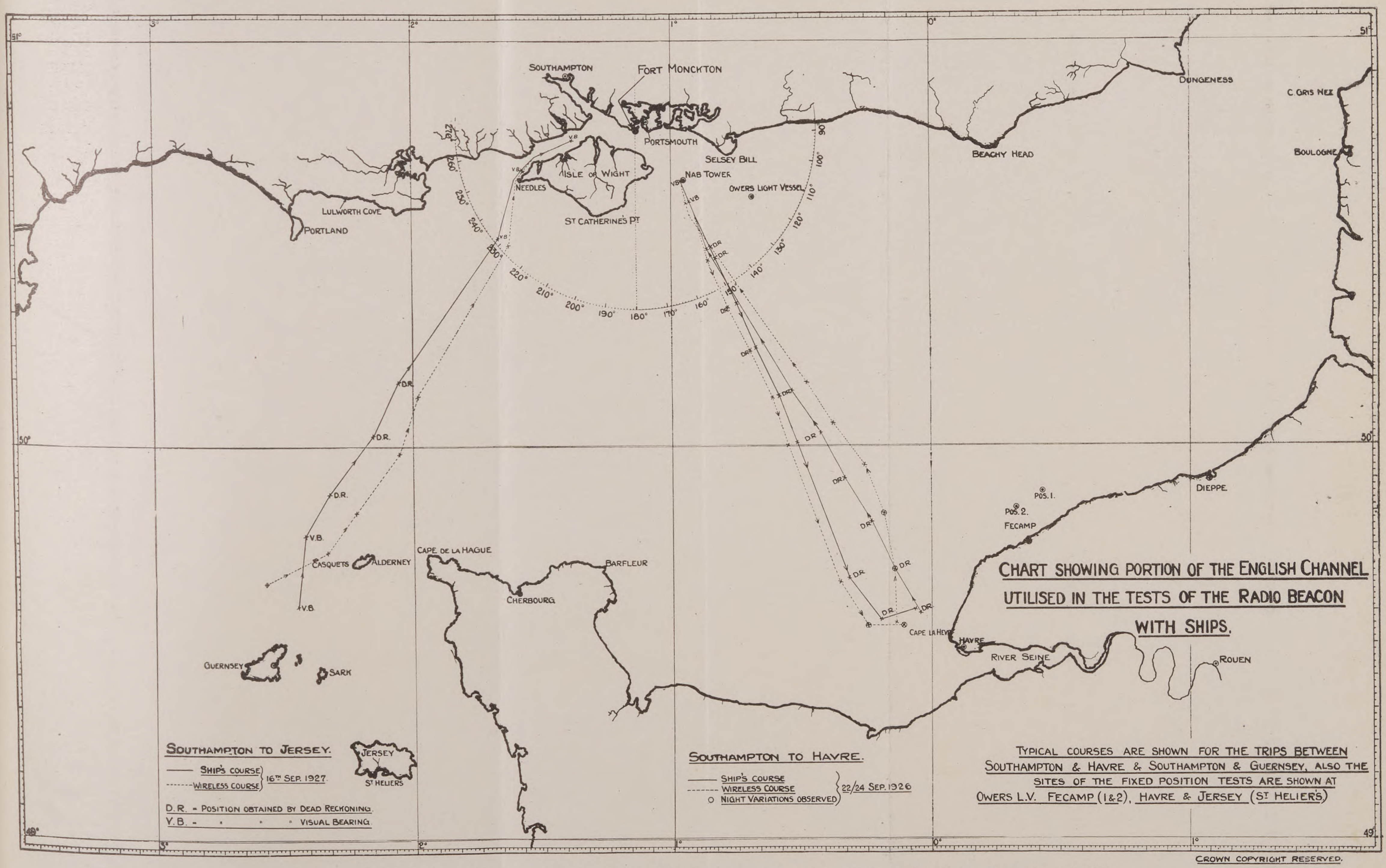

Fig. 13—Chart Showing Portion of the English Channel Utilized in the Tests of the Radio Beacon with Ships.

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servations can be largely mitigated by taking the mean of a number of successive bearings. It was not found possible to draw any definite distinction between the use of continuous wave and interrupted continuous wave transmissions.

Owing to the difficulty of anchoring a ship in a fixed position at a considerable distance from land for a period of 12 hours or more, it has not been possible to fill in the gap which occurs in these tests between distances of 23 and 92 miles, all oversea. No evidence of night effect was observed in the cross-channel tests at any distance below 60 miles, however, and it seems reasonable to infer therefore that the reliability of the bearings at 23 miles as recorded above would be maintained up to distances of 50 or 60 miles so long as the path of transmission is entirely oversea. Under these conditions we may expect that about 85 per cent of all observed bearings will be correct to within 2 deg., the remainder being subject to a maximum error of 6 deg. At larger distances night errors begin to be encountered but do not become serious until the oversea range is in excess of 90 miles.

This conclusion may be compared with the somewhat similar one contained in the following quotation from a paper[65] by the author, describing experiments made a few years ago to test the reliability of wireless direction-finding oversea.

> When, however, the transmission is entirely oversea, so as to be well clear of any land effects, the accuracy of observed bearings is maintained within the limit of 2 deg. for night as well as for day working at distances approaching 60 miles; and even up to 100 miles over 90 per cent of the bearings are correct within 2 deg., the limiting error under ordinary working conditions being 4 deg.

Such accuracies refer to the bearings obtained at a *land* direction-finding station; and are superior to those obtainable with a ship direction-finding installation, a matter which is discussed in Section 6 below.

Comparison of Bearing Observations by Beacon and by Direction-Finder

With a view to obtaining experience on the relative merits of the rotating beacon and the direction-finder as a means of obtaining wireless bearings, a direct comparison of the two systems has been carried out under a variety of conditions.

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At each site a simple receiver with a vertical aerial was employed for the reception of the signals from the rotating beacon. Adjacent to this, but at a sufficient distance to avoid mutual interference, a single-frame coil direction-finder was set up for the purpose of taking ordinary direction-finding bearings while the beacon loop was fixed with its plane approximately in the direction of the receiver. As a further precaution against producing a spurious reading on the direction-finding set the vertical aerial was disconnected during the taking of bearings on the direction-finder. A few tests carried out on land showed that apart from the occurrence of local errors due to the site selected, the two systems gave observed bearings of similar accuracy. In order to study the variations in bearings experienced with the two systems over extended periods, further experiments were carried out on a ship at Jersey and on land at Slough. The analysis of a large number of readings taken in this manner confirmed the similarity in accuracy of the two systems, and showed also that results obtained with continuous wave transmission were similar to those obtained with interrupted continuous waves.

At night time the variations in observed bearings were quite serious in all cases. The extent of the variation ranged from 20 deg. to 57 deg. in the case of the rotating beacon bearings, and from 69 deg. to 130 deg. in the case of the direction-finding bearings. Under these conditions, however, over 86 per cent of the bearings taken on the rotating beacon were in error by less than 10 deg. while in the case of the direction-finding bearings over 62 per cent were in error by less than 10 deg. As already mentioned it is now well-known that these errors are due to the reception of waves from the transmitter deflected in the upper atmosphere, and it is realized that in the results quoted above the variations on the direction-finder are exaggerated by the fact that the loop transmitter radiates very much more to the upper atmosphere than is the case with a vertical aerial.

Comparison Tests at Sea

Some of the test runs made on ships between Southampton and Jersey, discussed above, were carried out in a ship which is fitted with a Marconi direction-finder using fixed frame coils on the Bellini-Tosi system; and the opportunity was thus provided of comparing the two systems of obtaining wireless bearings

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under actual sea-going conditions. The results obtained on one trip, in which observations were made successively on the rotating beacon and on the direction-finder with the beacon fixed in the maximum signal position, are given in Table VIII.

Considering first the results in Table VIII it is seen that with two exceptions the direction-finding bearings agree with

TABLE VIII

Summary of Observations Obtained of a Return Trip between Southampton and Jersey, September 14 to 16, 1927

Date	G.M.T.	From information supplied by Captain of ship				Wireless Bearings:		Type of trans-mis-sion	Re-marks
		Position and how obtained		From Fort Monckton		from bea-con	by D.F.		
				Dis-tance	Bear-ing				
				miles	degs	degs.	degs.		
14.9.27	2340	Off Egypt Point.	V.B.	9.	90.	92.	88.	I.C.W.	
15.9.27	0010	Off Yarmouth Pier.	V.B.	16.	78.	74.	75.	C.W.	Sea rough
	0040	nr. Needles	V.B.	22.	68.	65.	67.	I.C.W.	
	0610	At Guernsey	A.	110.	34.	36.	33.	C.W.	
	0710	St. Martin's Pt.	D.R.	111.	31.	34.	33.	C.W.	
	0740	nr. Groznez Pt.	V.B.	114.	27.	30.	37.	I.C.W.	
	1540	At Jersey	A.	116.	21.	22.	26.	I.C.W.	Means of several readings
	1610	At Jersey	A.	116.	21.	22.	24.	C.W.	
	1640	At Jersey	A.	116.	21.	22.	25.	I.C.W.	
16.9.27	0710	nr. LaCorbière.	V.B.	119.	25.	28.	Sigs. too weak	C.W.	Sea smooth
	0940	nr. Casquets	V.B.	96.	35.	39.	36.	I.C.W.	
	1010	beyond Casquets	V.B.	88.	39.	36.	36.	C.W.	
	1040	49° 54′ N. 2° 17′ W.	D.R.	78.	40.	38.	42.	I.C.W.	
	1110	50° 2′ N. 2° 9.5′ W.	D.R.	68.	41.	38.	41.	C.W.	
	1140	50° 9.5′ N. 2° 2.5′ W.	D.R.	55.	43.	39.	41.	I.C.W.	
	1310	Outside Needles	V.B.	30.	52.	48.	57.	C.W.	
	1340	Inside Needles	V.B.	20.	71.	70.	71.	I.C.W.	
	1410	Lepe Middle	V.B.	14.	82.	82.	93.	C.W.	
	1440	Beacon Buoy.	V.B.	10.	119.	121.	119.	I.C.W.	

Notes.—(1) A = moored in dock, V.B. = visual bearing, D.R. = dead reckoning. (2) Maximum difference between estimated and observed bearings on beacon = 4 deg. in daylight. The observed bearings have been corrected for land errors in accordance with the previous calibration of the beacon in the area occupied by the ship. (3) Maximum difference between estimated and D.F. bearings = 11 deg. in daylight. Part of this difference is probably due to land effects for which a correction is not readily applicable.

the estimated bearings to within 5 deg. Of the two exceptions, one shows an error of 10 deg. at a range of 114 miles, while the second shows an error of 11 deg. at 14 miles distance, when the ship was proceeding up the Solent. It is thought that a portion of these errors, but probably not more than 3 deg. or 4 deg., is due to a land deviation effect which is difficult to compensate for

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on a ship direction-finding set. During the same run, the rotating beacon bearings, after correction for the 1 deg. or 2 deg. of error in some positions from the calibration curve, show a maximum departure from the estimated bearing of 4 deg.

In searching for an explanation of the direction-finding errors above, it is to be remembered that the direction-finding bearing is observed relative to the ship's head and that, therefore, the accuracy of the bearing is limited to the accuracy with which the ship's compass indicates the *instantaneous* direction of the ship's head. On the ship in question the direction-finding set is operated from the wireless room and on a signal being given to the bridge the compass reading is taken by one of the ship's officers. In anything but a very calm sea it is naturally difficult to keep the ship's head steady to within one or two degrees, and in a rough sea the swing may amount to five degrees or more. Also there is probably a lag between the actual direction of the ship's head and the reading of the compass. Finally, unless it is calibrated at frequent intervals, it is doubtful if the reading of a magnetic compass is accurate to within 2 deg. or 3 deg. over all parts of the scale. These considerations may result in the accuracy of a ship direction-finder installation being appreciably inferior to that of a similar apparatus on land. The errors due to the above cause will probably decrease with an increase in the size of the ship, but the ship mentioned above is probably typical in size to many which will, in the future, utilize wireless bearings for navigation. It is also to be remarked that the ship direction-finding set is subject to a quadrantal error, for which correction or compensation is necessary.

On the other hand the rotating beacon method of obtaining wireless bearings is free from all these objections. Except for the interfering effects of noise and general vibration, the accuracy of the observed bearings is the same whether the ship is in dock or at sea.

Reception of Beacon Transmissions on Loop and Aerial

Since it may sometimes be required to receive and observe signals from a rotating beacon on a closed loop or direction-finding set instead of on an open aerial, it was considered to be useful to make a brief comparison between the two cases. When working in daylight or under such conditions at night as to be

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free from night variations, it has been found that the bearings observed on a loop agree with a maximum departure of one degree, with those observed on an aerial. When operating under conditions of night variations, however, the errors encountered on the loop receiver are greater than those observed on the aerial. This effect is due to the fact that with downcoming waves arriving at the earth's surface the resultant horizontal magnetic field (operating on the loop) is of greater intensity than the vertical electric field (operating on the vertical aerial).[50,52] These results have been confirmed for both continuous wave and interrupted continuous wave transmission.

Night Errors Experienced Under Strictly Reversible Conditions

Reference has been made in one or two places in this paper to the application of the reciprocal theorem to wireless direction-finding. While the form of comparison of the rotating beacon and the direction-finder adopted in the tests described above is the correct one for determining the relative advantages of the two systems for marine navigation purposes, it does not constitute a strict reversibility of the system, aerial to loop. By erecting a direction-finder in close proximity to the rotating beacon and using a distant aerial either to transmit to the direction-finder or to receive from the beacon this reversibility is achieved. In this manner it was shown that both local and night errors are of the same order on the two systems and a graph of bearing taken at night under these conditions is shown in Fig. 14.

5. Conditions Affecting Direction-Finding as an Aid to Navigation

In the application of a direction-finding system to either marine or aerial navigation it is important to understand clearly the exact conditions under which the observed bearings are accurate, and in other conditions to appreciate the order of magnitude of the possible errors involved and the means of mitigating these where possible. It is generally conceded nowadays that an accuracy of 2 deg. in observed bearings is suitable for navigation purposes, particularly when it is borne in mind that the observations may be repeated at frequent intervals.

An important factor to be noted from a navigational point of view is that of any effect of fog on direction-finding, since it

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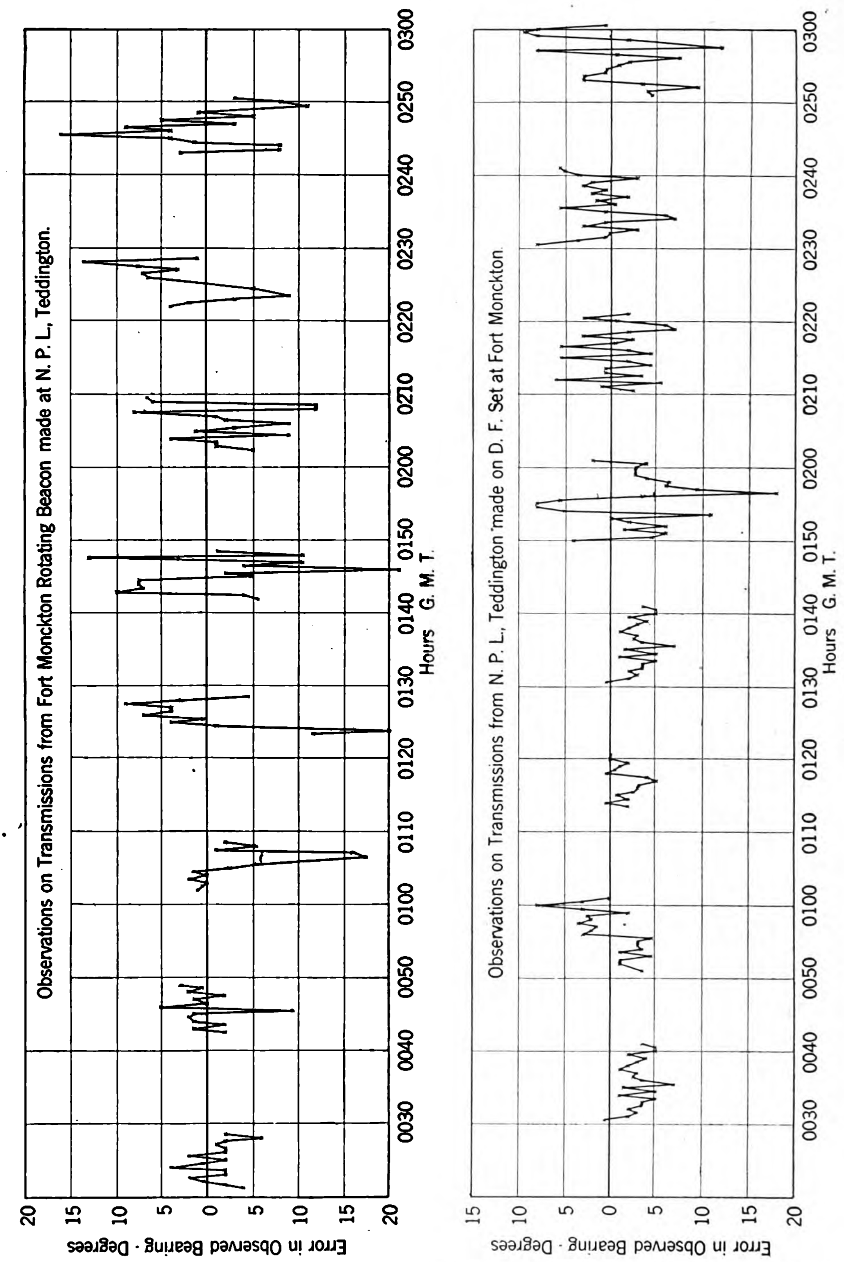

Fig. 14—Comparison of Night Observations on Rotating Beacon and on Direction Finder, March 22, 1928 (undamped waves, $\lambda = 525$ meters).

is chiefly during foggy weather that the majority of direction-finding stations are called into action. On several occasions the author has taken particular notice of the existence of fog at times when observations were in progress with a negative result. On one occasion in particular the fog was spread over the British Isle and a large portion of Western Europe, but the directional variations showed nothing beyond the usual day and night effects. These results form a confirmation to the observations of Rothé,[66] who concluded that no variations in ordinary atmospheric conditions would account for the small variations in bearings observed in the daytime.

During the course of the many experiments carried out with the direction-finder and the rotating beacon it has been found that the type of wave transmission employed has no effect upon the bearings observed, and that the errors and effects encountered are of the same order for damped, undamped, and interrupted undamped waves.[67,44]

(*a*) *Errors Due to Local Conditions*

In Sections 2 and 3 above it has been mentioned that various conditions may be present in the neighborhood of either a direction-finding station or a rotating beacon to cause an error in the bearing observed on a distant station. From various detailed investigations into the causes of such errors it is known that the most prominent effects are due to masses of metalwork and wires, either above or below the earth's surface, and to trees.[19,20] In most cases where the directional station is situated on land it is possible to select a site which is largely if not entirely immune from such effects, and provided that the residual errors do not exceed 2 deg. or 3 deg. in magnitude they can be definitely ascertained by a calibration at short distances, and a correction obtained for application to other observed bearings. In the case of the use of a direction-finder on board ship it is impossible to be clear of the metalwork of the ship itself and the resulting bearings are subject to a quadrantal error, which can either be compensated by circuit adjustment or corrected for from a chart.[41,42,43]

(*b*) *Coastal Errors*

Another type of error (which might be classified as one due to local conditions, but which, on account of its importance in

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the application of direction-finding to marine navigation deserves a special section) is that due to the deviation of wireless waves in crossing a coast-line, when the path of the waves lies approximately parallel to the coast. In the experimental work carried out at Orford, Suffolk, the coastal error on wavelengths of 450 and 600 meters was found to be of the order of 3 deg. or 4 deg. when the direction of transmission was within 20 deg. of the coast-line. In one instance, in which the wavelength was systematically increased from 500 to 2,600 meters, the corresponding error decreased from 3.2 deg. to 1.4 deg. On higher wavelengths the observed coastal error was less than one degree. In every instance the error was such as to indicate a bending of the waves towards the normal to the coast-line, in passing from the sea to the land side of the boundary. These experimental results are in complete agreement with those previously obtained by Eckersley[8] but, as already pointed out,[68] they are inconsistent with the explanation of the deviation as a coastal refraction effect due to the difference in the superficial velocity of wireless waves over land and sea. For the theoretical aspect of the problem would appear to indicate a deviation much smaller than that actually observed, and also in the opposite direction.

More recently some experiments on the propagation of waves across a coastal boundary have been made by Baümler and Zenneck.[69] Their results generally confirm those given above as to the magnitude and direction of the deviation of the waves in passing from sea to land. They further showed that the deviation at the coast is unaltered by the change in sea-level from high to low tide. Worlledge[70] has also given brief details of some experiments carried out at a direction-finding station in which indications were given of the reflection of wireless waves from the landward side of a coastal boundary. As a result of this phenomenon errors in observed bearing and indistinct minima due to interference may be produced at a direction-finding station when observing on transmissions emanating from a station on the same side of the boundary.

In connection with these coastal errors it is worthy of note that, with the accumulation of experience of the use of direction-finders on board ship, it is now becoming customary to mark out on charts the "arcs of good bearings" cf various transmitting stations, within which the results of observations are found to be reliable.[71] Similar arcs can, of course, be provided for shore

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direction-finding stations or rotating beacons; and their boundaries would be found in the initial calibration of the station for the permanent local errors.

(c) *Night Errors and Their Elimination*

As the result of the analysis of a large quantity of data obtained with wireless direction-finders it can be stated that the minimum distance at which night variations have been consistently experienced is about 30 miles for overland working, observations taken at closer ranges than this showing a maximum error of only 2 deg. or 3 deg. Mesny[58] has stated that night variations may occur for ranges of transmission as small as 15 miles, but according to the author's experience night errors are small at such distances. For example, observations of the bearings of the London broadcasting station taken at Slough, 18 miles away, have shown a maximum error of less than 5 deg. during many months (see Table IV). Similarly, the bearings of the Air Ministry in London, at Teddington, distance 17 miles, have shown a maximum error of 4.5 deg. while 99 per cent of the readings were within 2 deg. of the mean bearing. On the other hand, Ongar, which is 30 miles away, gave an error up to 8 deg. at the same observing station, and only 14 per cent of the readings differed from the mean by less than 2 deg. When, however, the path of transmission is entirely oversea, as was the case in the observations carried out at Orford on the transmissions from various ships, the above minimum distance is increased to about 80 or 90 miles.[65] This fact is of great importance in the application of direction-finding to marine navigation, and it is perhaps fortunate that the usual conditions connected therewith are that the path of transmission is entirely oversea and that the observed bearings should be accurate at distances which are usually less than 50 miles and never greater than 100 miles. It is now known that under such conditions direction-finding is sufficiently accurate for navigational purposes.

The explanation of the greater range freedom from directional errors due to the diminished attenuation of the direct wave travelling oversea was first given by Wright[73] in 1920. An exactly similar result has been obtained from the experiments carried out at sea in the observation of bearings from a rotating beacon.[44] Even in the presence of night errors their effect can be

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forming Fig. 15. It is generally assumed that the downcoming waves have travelled through the upper regions of the earth's atmosphere without deviating laterally from the great-circle plane through the transmitter and receiver. The author believes he is correct in stating that this assumption remains undisputed at the present time, except for wavelengths below about 50 meters. It will be evident, therefore, that any receiving system which is unaffected by a horizontal component of electric force would be free from night errors, even though the vertically polarized downcoming waves might still produce variations in received signal strength. A direction-finding receiving arrangement which fulfils this condition was patented by Adcock[14] in

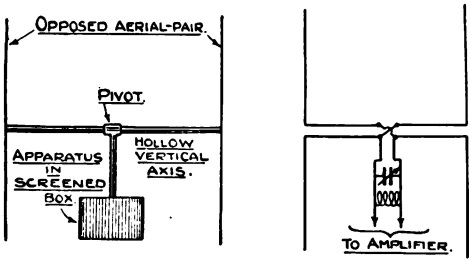

Fig. 16

1919, but this system does not appear to have received practical consideration until Mr. Barfield and the author[74] experimented with it in 1926.

The simplest form of the Adcock aerial system is a pair of spaced vertical aerials arranged to rotate about a central vertical axis, as shown in Fig. 16, thus forming the equivalent of the single closed-coil direction-finder. By making all connections to the centers of the aerials the horizontal members of the system are compensated so that no emf is induced in the system by a horizontal electric force. To obtain increased sensitivity larger aerials set at a greater spacing may be employed, and when the system becomes too large to rotate, two pairs of such aerials may be employed in conjunction with a radiogoniometer exactly as employed with closed loops in the Bellini-Tosi direction-finder. The schematic arrangement of such a system is shown in Fig. 17, while the photograph in Fig. 18 gives a view of the practical arrangement employed in a series of experiments which have been described in detail elsewhere.[74] The type of result obtained is illustrated by the curves given in Fig. 19, which shows the

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results of simultaneous observations made on an ordinary closed-coil direction-finder and on one of the Adcock type on the transmissions from a British broadcasting station. It will be seen from this diagram that while the closed-coil set during this period of observations was giving very bad night variations, the corresponding errors observed on the Adcock direction-finder were negligible. It is to be concluded from these and similar results obtained on the transmissions from other broadcasting stations that lateral deviation plays a negligible part in producing the large and variable errors which are obtained at night on the present type of closed-coil direction-finding set, and that therefore these errors are almost entirely caused by the arrival of downcoming waves polarized with the electric force horizontal.

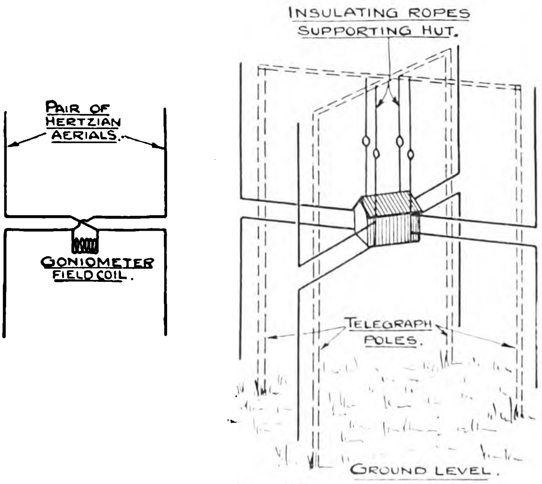

Fig. 17

It will be appreciated that it follows from the above conclusions that such a system may be used as a direction-finder which gives the true great-circle plane of arrival of wireless waves, whatever may be their state of polarization or their angle of incidence at the earth's surface. The system should, therefore, have important applications as an accurate direction-finder under night conditions, or for use in observing on transmissions from aircraft at high angles of elevation, when the ordinary closed loop type of instrument is subject to large errors. Conversely the Adcock aerial system may be used in place of the closed loop at a directional transmitting station for use either as

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a course-setter or as a bearing indicator, with freedom from the night errors which have already been experienced.[33,44] The practical development of the system both as a transmitter and as a receiver towards the above ends is now being pursued in Great Britain.

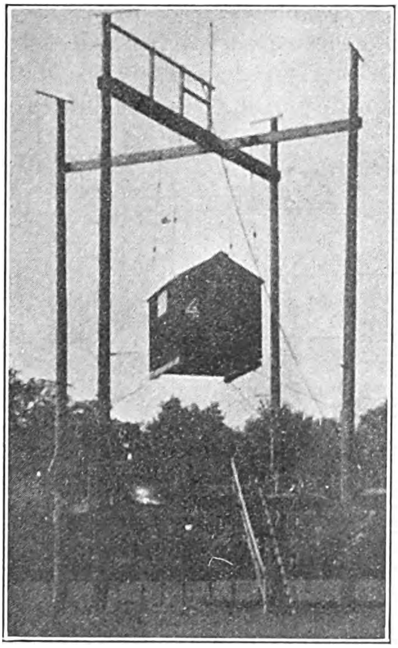

Fig. 18—View of Hut Suspended at the Center of an Adcock Aerial System. The hut contains the radiogoniometer and receiver used in taking bearings free from night errors.

6. Relative Advantages of Direction-Finding by Transmission and Reception for Navigation Purposes

(a) *Location of the Direction-Finder*

At the outset it must be admitted frankly that, owing to the more advanced state of its development at the present time, the wireless direction-finder when used in a fixed position on land gives a somewhat superior accuracy to the rotating beacon; for it is not easy to obtain bearings on the rotating beacon to a greater accuracy than two degrees, whereas a good land direction-finding station should give bearings reliable to one degree. This,

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however, is not the whole story and reference must be made to the conclusions of a controversy of long standing on the matters of the location of the direction-finder, whether on ship or shore, and of the individual upon whom it is most desirable to place the responsibility of observing the bearings. Without giving the full details of this controversy it may be stated that it is very desirable for the navigator to take the wireless bearings personally or have these observations made by somebody under his direct control and supervision. Since the navigator is responsible for the safety of his ship it is unreasonable to expect

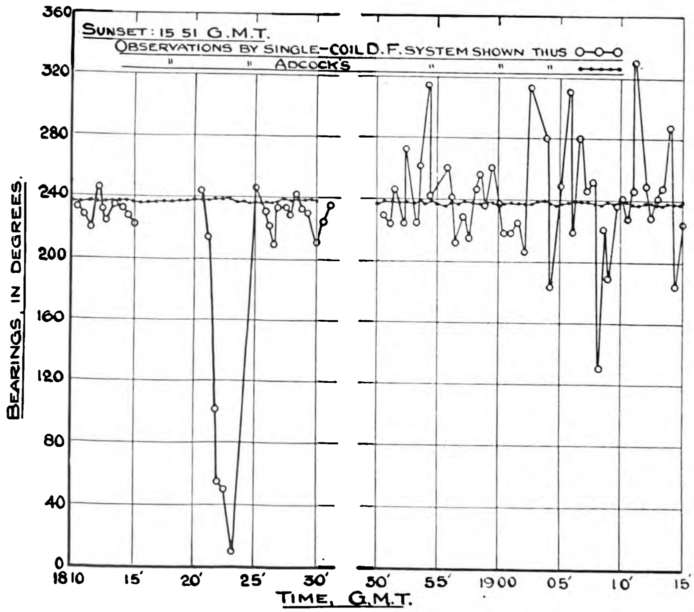

Fig. 19—Observations of Bearings on Bournemouth, December 10, 1925 ($\lambda = 386$ m)

him to rely, particularly in emergencies, upon observations made at a shore direction-finding station by a man quite unknown to him, and of whose skill and reliability he is quite unable to judge. In addition there is the possibility of error in signalling the bearings from shore to the ship, the interference to other wireless traffic in doing so, and the point of some strategic importance that the information as to the ship's bearings or position is broadcast to all wireless stations in the vicinity. It will be evident, therefore, that the most desirable position for the wire-

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less direction-finder is on the ship itself, and this view is confirmed by the fact that during the past few years an increasing number of ships have been fitted with direction-finders whereas the number of shore direction-finding stations has never been large and is decreasing. The true comparison to be made, therefore, is between the rotating beacon transmitter set up on shore and the direction-finder installed on board ship. In the following paragraphs the discussion of this comparison is classified according to the superiority which one system appears to have over the other.

(*b*) *Superiority of the Ship Direction-Finder over the Rotating Beacon*

It was considered at one time that with the installation of the direction-finder on board ship the necessity for special transmissions for wireless bearing purposes would be eliminated. This, however, has been found not to be the case and it is now the practice to erect fixed transmitting stations or beacons[75] for the sole purpose of providing special transmissions for the use of ship direction-finding sets. These beacons are usually located near lighthouses or on light-vessels at points of importance for navigation. The use of the direction-finder is not limited to such beacons, for bearings may be taken upon the transmissions from any land station the position of which is accurately known. It has been found, for example, that some broadcasting stations form very useful fixed beacons operating for many hours of the day when it is desired to take wireless bearings on a ship at sea. The navigator with a direction-finder under his control thus has the opportunity of taking a series of cross-check bearings and so of improving the accuracy of determination of his position.

A further advantage of the ship fitted with the direction-finding set is that the opportunity is provided for observing bearings upon the transmissions from other ships. This may be useful when contact is made with a ship which knows its own position, but it is of greater importance when applied to the location of ships in distress, as has already been demonstrated in several practical cases. Moreover, there is the possibility that the future development of the ship direction-finder will provide a means for reducing or avoiding collisions between ships in foggy weather.

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These advantages are not possessed by the rotating beacon system which provides the ships with the opportunities of obtaining bearings only during the operation of such beacons, and in those areas of the world in which these beacons are installed. The navigator with a direction-finding set under his immediate control would probably use it very frequently and would thus soon become aware of the occurrence of any fault or error. With the rotating beacon system he would be dependent upon information from the shore station staff as to any such errors; but against this it may be urged that such a staff would be more expert at dealing with and removing the error than would the average navigator or wireless operator.

(c) *Superiority of the Rotating Beacon over the Ship Direction-Finder*

By way of countering the remarks made in the above section it may be said that the rotating beacon transmitter retains all the advantages which were sacrificed when the direction-finding installation was moved from the shore to the ship. These advantages comprise the fact that the directional part of the system is fixed in position on solid ground and that it can, therefore, be accurately oriented and calibrated at installation, and also that it can be under the continuous supervision of experts. One directional shore station will provide service to an almost unlimited number of ships, whereas each ship direction-finder must be carefully installed and operated by experienced men.

One of the great advantages of the rotating beacon system is that no apparatus beyond a simple wireless receiver and a chronograph is necessary on the ship itself; a direction-finding service is thus provided for ships of all classes from the largest to the smallest. In the case of many small ships it would not be practicable to install a direction-finder with any pretense to accuracy; and to the large ship already fitted with a direction-finding set a rotating beacon would be an additional asset in enabling further bearings to be taken either for check purposes or for position determination. Furthermore, the method of observing bearings is so simple that they can easily be determined by the navigator himself in any portion of the ship without necessarily requiring the assistance of a wireless operator.

It has already been hinted that the accuracy of bearing observation by direction-finder on a ship at sea is less than that

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obtainable with a similar installation on land. The difference is, in fact, sufficient to make the optimum accuracy of the ship direction-finder comparable with that of the rotating beacon system. The direction-finding bearing is taken relative to the direction of the ship's head, and its accuracy depends upon the steadiness of the ship and also upon the accuracy with which the direction of the ship's head is given by the compass reading at any desired instant. In anything but a calm sea it is difficult to keep the head of a small ship steady to within one or two degrees, and in a rough sea the swing may amount to five degrees or more. Also there is probably a lag between the actual direction of the ship's head and the reading of the compass, whereas no such lag exists on the direction-finder. The bearing obtained from the rotating beacon is entirely free from these limitations and its accuracy is practically the same whether the ship is at sea or in dock. Also since the accuracy of observed bearing has been shown to be largely immune from conditions local to the receiver, no correction or compensation is necessary corresponding to the quadrantal error associated with the ship direction-finder. This error is likely to be serious in the case of many small ships, and there is the added possibility of its alteration with draught and nature of cargo.

The limitation of range of accurate bearings due to night effect has been shown, both theoretically and experimentally, to affect both systems of directional wireless to the same degree. The accuracy of the rotating beacon has also been found to be sensibly the same for both continuous wave and interrupted continuous wave transmission. Considering the possibility of interference of the rotating beacon transmitters with other wireless services, a similar objection may be raised to the fixed beacon transmitter which has been found to be so necessary in conjunction with the ship direction-finder. In neither case, however, is any serious interference likely to be caused since definite and distinct wavelengths have now been allocated to both types of beacon.

A small point just worthy of mention in concluding this section is that rotating beacons erected on the coast would be of some considerable service to aircraft navigators, particularly those making daily flights along the principal air routes.

(*d*) *Conclusions*

It will be gathered from the above discussion that the rotating beacon system is considered to have some advantages

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over the ship's direction-finder in the application of directional wireless to marine navigation. These advantages are, however, probably not sufficiently well defined at present as to make the future policy of those concerned with these matters independent of certain other factors; and in any case it is evident that the economics of the situation must be considered as well as the technical features. The rotating beacon system would at first appear to be an economy in that the whole cost of the directional apparatus is removed to the authority controlling the shore stations; and it is to be presumed that the cost of installing and maintaining the rotating beacons would be met by the levy of a tax upon all ships using the beacons. This might be a great advantage to the shipowners who, judging from their tardiness in equipping their ships with direction-finders, do not appear to find this apparatus very cheap. To the cost of such direction-finders, moreover, is to be added that of the fixed beacons specially built for their use. This cost depends upon local conditions and whether or not the beacon can be installed at an existing lighthouse and maintained by the personnel of the lighthouse. The first rotating beacon for practical working to ships has yet to be built so that details of the cost are at present unknown.

7. Acknowledgments

This communication is published by permission of the Department of Scientific and Industrial Research, London, England. The whole of the work described in this paper has been carried out for the Radio Research Board of that Department, and full acknowledgment must be made of the assistance rendered by the author's associates working for the Board in the various phases of the whole investigation.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

SOME EXPERIMENTS IN SHORT DISTANCE SHORT-WAVE RADIO TRANSMISSION*

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Summary—*Some experiments in short-wave radio transmission over a distance of 55 miles are described, the results of which are interpreted to indicate the presence of strong "sky" waves, with "ground" waves of negligible amplitude in comparison with the "sky" waves as received. Upon decreasing the transmitter wavelength, at a given time of day, a minimum wavelength was reached below which no communication could be obtained; this wavelength is termed the "cut-off" wavelength. The average value of the cut-off wavelength, for various times of day, is given for several different months. The minimum observed wavelength upon which communication was possible was 28 meters.*

A series of experiments in which an orientable half-wavelength antenna was employed served to indicate definitely an optimum position of the antenna for transmission over the 55-mile distance. The indicated transmission path left the transmitter at an angle of approximately 65 degrees to the horizontal. In long distance communication the position of the antenna was found to have no appreciable effect.

THE work herein described was carried out over a period of two years, over a fixed distance of approximately 55 miles between Round Hill, South Dartmouth, Mass. and Auburndale, Mass. on wavelengths ranging from 25 to 80 meters. In view of the results obtained which are explainable in a rather straight-forward manner on the basis of refraction of the radio waves in the upper atmosphere, the presence of such a refracting medium will be assumed as an hypothesis. It was early found in the course of the experiments that at times no communication could be obtained on the wavelengths, and at the time of day, when under normal conditions very strong signals would be received. This phenomenon is believed to be due to conditions of the refracting layer such that it was impossible for refracted waves to return to the surface of the earth within the fixed distance of 55 miles. The noticeable absence of signals during these periods indicated that the "ground" signals, due to waves

* Dewey decimal classification: R113. Original manuscript received by the Institute, November 7, 1928. Contribution from the Round Hill Short-wave Radio Research, sponsored by Col. E. H. R. Green.

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following the surface of the earth, did not play any appreciable part in contributing to the usual received signal.

A very complete theoretical study of the transmissions paths of short waves, through an electron atmosphere, has been made by Baker and Rice.[1] Their conclusions are here utilized in brief form, in the summary of the following paragraphs. For further interesting and valuable discussions of the problem the reader is referred to the books and papers below.[2]

Fig. 1 is a sketch showing the paths of the various rays radiated by an ideal transmitter having a characteristic as sketched

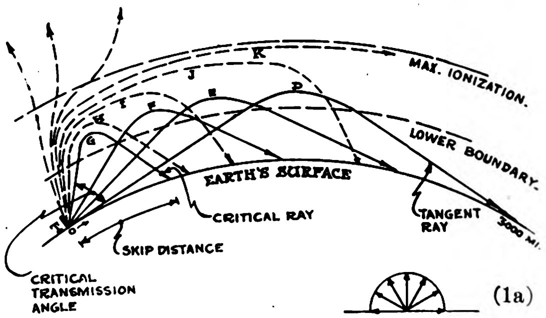

Fig. 1—Sketch Showing the Paths of the Various Rays Radiated by an Ideal Transmitter Having a Characteristic as Sketched in Fig. 1a, as Given by Baker and Rice.

in Fig. 1a, as given by Baker and Rice. The tangent ray, *TD*, has the greatest range of all the rays radiated close to the horizontal. The critical ray, *TG*, traverses the minimum distance along the surface of the earth. Rays leaving the transmitter near the vertical may return to the surface of the earth at great distances from the transmitter, as *TJ*, *TK*. Rays which penetrate to the region of maximum ionization are given curvatures

[1] W. G. Baker and C. W. Rice, "Refraction of Short Waves in the Upper Atmosphere," *Jour.*, A.I.E.E., **45**, p. 535–539, June, 1926; also reprints, of a more detailed nature, of paper presented to A.I.E.E. in February, 1926.

[2] A. H. Taylor, "Relations between Height of Kennelly-Heaviside Layer and High Frequency Radio Transmission Phenomena," Proc. I. R. E., **14**, 521–540; August, 1926. A. H. Taylor and E. O. Hulburt, "The Propagation of Radio Waves over the Earth," *Phys. Rev.*, **27**, February, 1926. Also *QST*, October, 1925. H. W. Nichols and J. C. Schelleng, "Propagation of Electric Waves over the Earth," *Bell Sys. Tech. Jour.*, VI–**2**, 215–235, April, 1925. P. O. Pederson, "The Propagation of Radio Waves," published by "Danmarks Naturvidenskabelige Samfund," 1927, and sold by G. E. C. Gad, Vimmelskaftet 32, Copenhagen, K. Denmark. (In English.)

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departing only slightly from that of the layer itself; the rays may consequently pass around the earth before returning to the surface or may be bent out into space. Rays emitted at angles very near to the vertical pass through the layer and out into space with relatively small changes in direction.

In Fig. 1 an ideal transmitter is assumed at *T*, i.e., a system which radiates equally well in all directions above the surface of the earth— as sketched in Fig.1A. Considering now the range of rays emitted at various angles above the horizontal, we find that the tangent ray *TD*, if unaffected by surface conditions, will return to the surface of the earth at a distance, in round numbers, of 3,000 miles from the transmitter *T*. This distance is only slightly dependent upon the wavelength and the conditions of the refracting layer, because of the relatively small portion of the path which is traversed within the refracting layer. For increasing angles of transmission, the range is decreased, very rapidly at first and then more and more slowly until the minimum range is reached (*TG*), at which point the change of range as a function of transmission angle is zero. The path for the minimum range is very critically influenced by the conditions of the refracting layer (height, thickness, value of maximum ionization density and the distribution of ionization), because of the greater penetration and the greater curvature given the ray, and consequently the path varies rapidly with the wavelength. If higher angles of transmission are considered, the range increases as the angle is increased, slowly at first and then more and more rapidly. At the higher angles, the rays might travel around the earth before being given a sufficient curvature to return them to the earth's surface. For angles near the vertical, the rays are bent but little, and pass through the layer and out into space.

The particular point of interest in the summary above is that, with fixed layer conditions, there exists a minimum range for the refracted rays, associated with a definite transmission angle. This minimum range is many times confused with the "skip-distance," i.e., that range over which no signals at all are received. When quite short waves are used, so that the minimum range is of the order of several hundred miles, this confusion is not serious, because of the usually limited range of the ground wave. For short distance operation, however, the ground wave intensity becomes very important in determining the actual skip-distance. The observed skip-distance, in these cases, depends not only on

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the minimum range of the refracted waves (which defines the outer skip-distance boundary) but also on the maximum range of the ground waves (which defines the inner skip-distance boundary). Any factors tending to decrease the first, or to increase the second, will result in a decreased skip-distance. If the ground waves reach the minimum refracted wave range with an appreciable intensity, then no skip-distance is encountered, as, for example, in short-wave transmission from aircraft. If the surface waves are highly attenuated near the transmitter, then the only recourse for short distance short-wave communication is to utilize relatively high angle rays.

For certain layer conditions, short distance communication will not be possible unless the power of the transmitter is increased to a point where the attenuation of the ground waves is overcome by sheer force to provide a signal of practical intensity. In these experiments the power output of the transmitter was purposely restricted to 50 watts, or less, in order that good reception would be attained only when favorable transmission conditions existed, making the delineation between favorable and unfavorable conditions as marked as possible. In practice there is little difficulty in identifying the "ground" wave and "sky" wave signals as received. The former entirely lack the general characteristics of short-wave transmission, i.e., violent and rapid changes in amplitude (even when the transmitter is carefully controlled as regards frequency and amplitude of the antenna current); the ground wave signals are remarkably steady, reminding one of very long wave signals.

For long range communication, 5,000 miles or more, it is seen from the considerations above that two classes of transmission path are possible. The first employs the rays radiated close to the horizontal at the transmitter, which are effective to distances of approximately 3,000 miles. To cover ranges greater than this it is necessary to invoke the aid of multiple reflections from the earth's surface. The idea is theoretically tenable, but has no great appeal from a practical viewpoint owing to the high absorption encountered by the waves at surface grazing regions, and the scattering effect of surface irregularities. The second class of path is provided by any rays which cover a range greater than 3,000 miles, invoking no multiple reflections. These paths are the high angle paths, indicated by the dashed curves of Fig. 1. The loss of energy due to dispersion of these paths over the

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earth's surface may in a large measure be overcome by the low attenuation, resulting in signal amplitudes of practical interest.

The experiments of Meissner[3] with a "searchlight" beam transmitter definitely indicate the usefulness of comparatively high angle radiation in effecting communication between Germany and Argentina on a wavelength of 11 meters. Two transmission angles for maximum received signal strength were found within the range of the observations, one at approximately 80 degrees and the other at approximately 37 degrees with the horizontal. It is regrettable that the physical limitations of the reflectors did not permit of swinging the beam to the horizontal, instead of a minimum of 30 degrees. Transmission at high angles

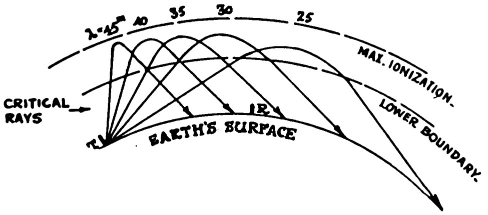

Fig. 2—Paths of the Critical Rays (*TG* of Fig. 1) for Various Wavelengths, Based on the Work of Baker and Rice.

over great distances is further substantiated by the general experiences of amateurs, using low power, and, in the majority of cases, essentially horizontal antennas at such heights above earth that a strong component of radiation is sent out at high angles.

Returning to the specific problem of short-distance communication, where the distance to be covered was sometimes less and sometimes more than the minimum refracted ray range, observations were made to determine the minimum wavelength which would provide reliable reception at various times of day and in various seasons. With given conditions of the refracting layer, the minimum range of transmission will increase as the wavelength is decreased, somewhat as sketched in Fig. 2. For a given distance between transmitter and receiver, *T–R*, a wavelength of 40 meters would produce strong signals, while a wavelength of 35 meters would fail entirely. For some wavelength,

[3] A. Meissner, "Directive Radiation with Horizontal Antennas," Proc. I. R. E., 15, 930–934; November, 1927.

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here about 38 meters, the path of the minimum range ray would strike the earth in the immediate vicinity of the receiver. Any slight irregularities of the layer conditions would then cause the foot of the path to sweep back and forth across the receiving point, producing violent "fading" or "fluttering." Such was found to be the case in these tests, the signals swinging from relatively large amplitudes to zero. Under more stable conditions of the layer, the fluttering was not pronounced, and, under such conditions, the wavelength at which communication just failed or was "cut off," could be determined within about one per cent.

Fig. 2 shows the paths of the critical rays (*TG* of Fig. 1) for various wavelengths, based on the work of Baker and Rice. For each wavelength the distance from the transmitter *T* to the point at which the ray returns to the earth's surface is approximately the "skip-distance." For the conditions of the diagram, wave-

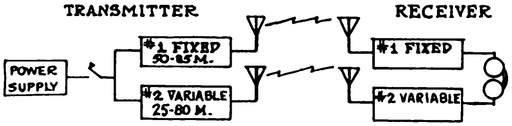

Fig. 3—Arrangement of Duplex Transmitter-Receiver Equipment Used in Making the Observations Described.

lengths longer than about 38 meters will be received at *R*, but wavelengths shorter than this value will not be received.

The equipment was arranged as in Fig. 3. Channel No. 1 was arranged as a communication channel, or "order wire," to be operative on wavelengths which had been found to give reliable communication at the time of day at which the experiments were being conducted. Channel No. 2 was arranged with a special transmitter designed for rapid manual change of wavelength. The transmitters were keyed simultaneously on the same power supply. The receivers were of similar design, consisting of an autodyne detector followed by two stages of audio-frequency amplification. Each receiver connected to one telephone receiver of a "split" headset.[4] If equal signal intensities were obtained on both wavelengths, and if the operator adjusted the autodynes to produce the same beat tone, the signal heard in the headset was remarkably similar to that obtained with a single receiver, except that the signal appeared to change from one ear to the

[4] J. K. Clapp, "Multiplex Short Wave Reception," *QST*, March, 1926.

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other, as the fading periods on the two wavelengths were not the same. No definite relationship between the fading periods on any two wavelengths was observed, even when the difference in wavelength was small.[5] Ordinarily the fading period, on the wavelengths covered in this investigation, is so short that the effect is that of an irregular amplitude modulation of the signal

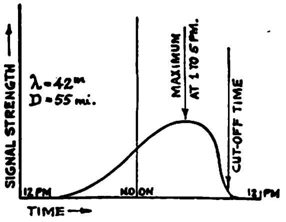

Fig. 4—Illustrating Average Variation in Received Signal Strength, for Short Distance Communication, as a Function of the Time of Day.

as received, made evident to the ear by a roughening of the beat tone without much variation of the average beat tone intensity.

By changing the wavelength of transmitter No. 2 in successive steps, a general idea of the "cut-off" wavelength at a given time of day and under given conditions of the refracting layer could readily be obtained. The received signal was found to vary with

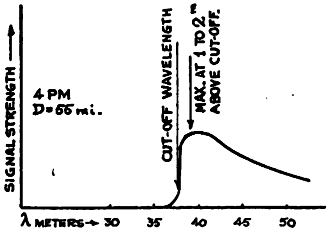

Fig. 5—Illustrating the Variation in Received Signal Strength, for Short Distance Communication, at a Given Time of Day, as a Function of the Wavelength of the Transmitter.

the wavelength as sketched in Fig. 5. When the signal had been reduced to a point where it could just be identified, though far too weak for communication, communication was considered as being "cut-off." Reception varied with the time of day, on a

[5] R. Brown, De L. K. Martin, and R. K. Potter, "Some Studies in Radio Broadcast Transmission," Proc. I. R. E., 14, 57–131; February, 1926.

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fixed wavelength, as sketched in Fig. 4. The values shown in Figs. 4 and 5 are for summer conditions. A decided drop in signal strength accompanied any marked increase in wavelength above the cut-off value. It was soon found, in making observations, that the value of the cut-off wavelength was subject to large and rapid fluctuations.[6] As interest was here centered on *average* values, no attempt was made to develop means for making more rapid determinations. The oscillographic method described by Heising could well be adapted to this study, as well as the method of "wobbling" the transmitter frequency.[7]

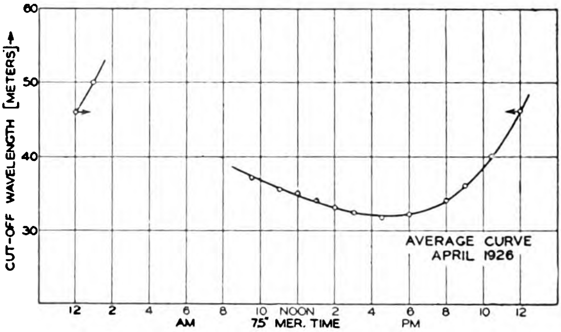

Fig. 6

The results obtained, over the fixed distance of 55 miles, are set forth in Figs. 6 to 11 inclusive. The early summer and mid-summer records show a rather flat curve, while the late fall records show a more or less regular steep curve. The cut-off wavelengths are plotted against the time of observation, all times being given in Eastern Standard Time. Signals on wavelengths lying just below the curve were not received, while, in general, good reception was obtained on wavelengths lying just above the curve; that is, on wavelengths just above the cut-off wavelength. Observations were made on three dates in April, 1926, between the hours of 9 A.M. and 1 A.M. of the succeeding day. The curve of Fig. 6 shows the minimum wavelength on which communication was possible in this period to have been 32 meters, occurring at about 5 P.M.

[6] R. A. Heising, "Experiments and Observations Concerning the Ionized Regions of the Atmosphere," PROC. I. R. E., 16, 75–99; January, 1928.
[7] R. A. Heising, J. C. Schelleng, and G. C. Southworth, "Some Measurements of Short Wave Transmission," PROC. I. R. E., **14**, 613; October, 1926.

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Fig. 7 indicates the results of tests during May, 1926, in which efforts were made to obtain data throughout the twenty-four hours. From 9 A.M. to 1 A.M. definite results were obtained, but from 1 A.M. until 9 A.M., particularly between 5 A.M. and 9 A.M.,

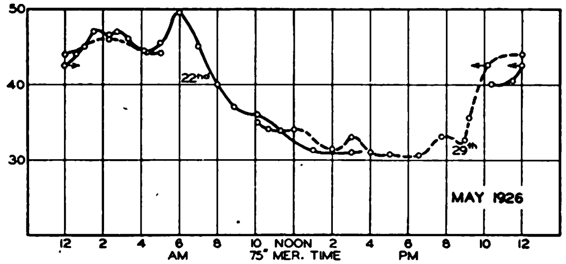

Fig. 7

the results were quite frequently indefinite. In every case the value of the cut-off wavelength rose abruptly just before sunrise, reaching a maximum shortly after sunrise and then gradually diminishing throughout the morning. The minimum wavelength on which satisfactory communication could be obtained during this period was 31 meters, attained at approximately 5 P.M.

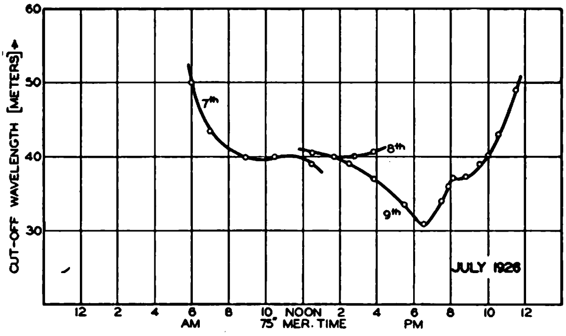

Fig. 8

Fig. 8 shows the results of a three-day test in July, 1926. During this period the signals obtained after midnight did not show any evidence of being due to refracted waves, but appeared to be due solely to the ground wave, as in this interval the signals were characterized by extreme constancy of amplitude. In the interval from 6 A.M. to midnight the results were quite definite, as shown by the curve. The minimum wavelength for communication in this interval was about 32 meters.

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At the end of July, 1926, a four-day test was made, after improvements in the methods of handling the tests had made possible more rapid determinations. The four curves of Fig. 9 show the great fluctuations in the value of cut-off wavelength which were encountered in relatively short intervals of time. The

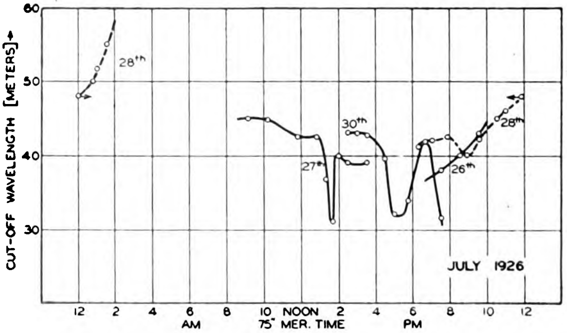

Fig. 9

average cut-off wavelength during these tests was higher than that obtained earlier in the month, but at times wavelengths very near to 30 meters were observed. The erratic nature of these curves is typical of the observations by the more rapid method, but it is surprising what a regular curve is obtained on averaging a large number of such observations.

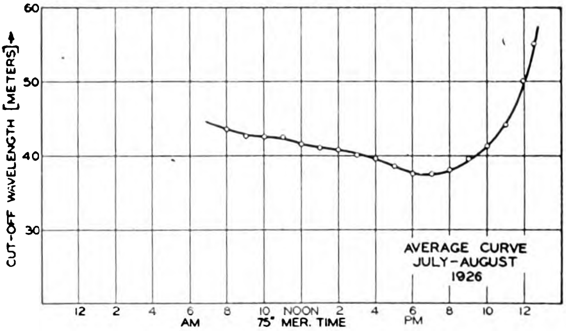

Fig. 10

All of the data for the tests of July and August, 1926, were averaged and the curve of Fig. 10 shows the result. The minimum wavelength of the average curve is 37.5 meters, where under temporary and more favorable conditions wavelengths as short

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as 31 meters were observed. The results of this averaging are so striking as to suggest strongly that the *average* conditions of the refracting layer undergo a regular variation. From time to time, "bumps" or "hollows," or regions of greater or less density of ionization, drift across the path of transmission, causing the transient irregularities which were observed.

Fig. 11 shows the average of the observations for October, 1926, the minimum wavelength observed being 29 meters at approximately 1 P.M. The curve as a whole has steepened materially, indicating that the number of hours of the day during which satisfactory communication could be obtained was decidedly less than in the preceding months. It would be expected

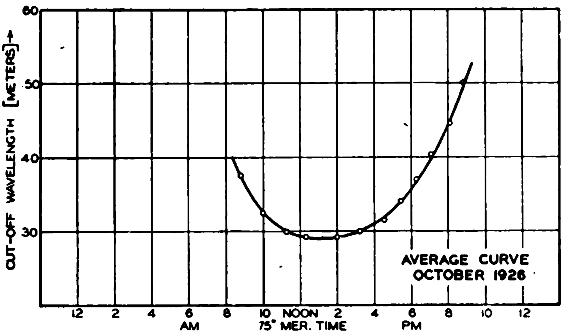

Fig. 11

that the minimum value of cut-off wavelength would increase as winter came on, owing to the reduced density of ionization, or to an increased average height of the refracting layer. The reduction in cut-off wavelength here encountered may be associated with the heavily overcast weather which was encountered during the period of the observations.

In this connection it was repeatedly observed that when large, well-defined clouds drifted across the path of transmission the received signals underwent a definite change in intensity. When the clouds were sufficiently well separated so that distinct observations could be made, it was found that the signal intensity rose when the cloud was in the line of transmission. On days which were heavily overcast, with or without fog at Round Hill (the station there being located within twenty yards of the shore of Buzzard's Bay), the average signal intensity was materially greater than on clear days. When rain was falling over the

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territory including both transmitter and receiver, the signals fell below the clear weather average.[8]

At times when the cut-off wavelength changed very rapidly, as between 7 and 9:30 P.M. during October, 1926, it was impossible to change wavelength with sufficient rapidity by manual methods. Resort was then had to the use of two transmitters at each point, the wavelengths being "staggered." For example:

	Round Hill		Auburndale
No. 1	35 meters	No. 1	37 meters
No. 2	39 meters	No. 2	41 meters

As the cut-off wavelength rose toward 35 meters, the signals from Round Hill would begin to flutter and then die out. Auburndale then communicated with Round Hill on 37 meters, until the cut-off wavelength rose to that value. Round Hill, on the extinction of their 35-meter signals, would transmit on 39 meters, thereby maintaining communication. If time permitted, Round Hill would increase the wavelength of transmitter No. 1 to some value higher than 41 meters, thus extending the range of the observations. By this means it was possible to obtain the times at which each of the four, or more, wavelengths were cut off. In many instances it was found that the cut-off wavelength rose 10 to 20 meters above its initial value in a period of ten or fifteen minutes. In the early evening the cut-off wavelength often reached the vicinity of 55 meters (above which no sharply defined cut-off effects were observed) from a value of 38 to 42 meters within fifteen minutes. Communication then usually failed entirely, except for faint signals on wavelengths of 80 meters or more, until the following forenoon. When communication failed abruptly in this manner, no instance was observed in which communication could be re-established on the same wavelength during the same evening. It appeared as though the characteristics of the refracting layer altered abruptly and did not return to the average normal until the changes of the ensuing day had taken place.

The above discussion of short distance transmission, including the interpretation on the basis on high angle refracted waves, is based on the theoretical premise of the work of Baker and Rice. In order to obtain an experimental verification of this viewpoint,

[8] "The Effect of Water Vapor in the Atmosphere on the Propagation of Electro-magnetic Waves," *Jahrb. d. Draht. Tel.*, **12**, 184; August, 1927. Also **19**, 58–72; January, 1922.

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a series of tests was conducted with an orientable transmitting antenna system.[9] The antenna proper consisted of fifty feet of copper tubing mounted on a light wooden lattice frame, carrying the transmitter at the midpoint. The center of the frame was supported on a universal joint at the top of a fifty-foot telegraph pole, set "in the clear" on the beach at Round Hill. Thus it was possible to place the antenna in any desired direction in space, the center of the antenna system being fifty feet above the surface of the earth.

The first series of tests was made to determine the best position of the antenna for transmission from Round Hill to Auburndale, using a wavelength of 42 meters. Definite results were very difficult to obtain because of fast fading; by graphically recording the received signals for a period of three minutes for each chosen position of the antenna, taking the average value of the record by planimeter, and taking a large number of observations, some indication of the average received signal intensity was obtained. An optimum position of the antenna was clearly indicated. The antenna was contained in a vertical plane passing through the transmitter and receiver, with the end of the antenna nearest to the receiver depressed some 25 deg. below the horizontal, as indicated in Fig. 12. Application of antenna theory to this configuration indicates that the antenna radiates with greatest intensity at an angle of about 65 deg. above the horizontal, in the direction of the receiver, and that the image of the antenna does not contribute materially to this radiation. The optimum position of the antenna indicates that the effective transmission path left the transmitter at an angle of 65 deg. to the horizontal, bearing out the idea of high angle transmission.

Consideration of the antenna and its image would lead one to expect fairly good transmission if the antenna were horizontal and at right angles to the line between stations; this was found to be true. If the antenna were placed in the line of the transmission path, then the only received signal would be due to the image, which would of course depend largely upon the character of the earth. With highly comducting earth, good signals should be obtained, while with highly absorbing earth the signals would be weak. In a few observations moderately good signals were obtained; in a very large number the signals were quite weak and

[9] G. W. Pickard, "The Polarization of Radio Waves," Proc. I. R. E., **14**, 205–212; April, 1926, and references given therein.

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in three instances no signals at all could be obtained. The experimental evidence is far from complete, but it indicates a condition of very imperfect reflection from the earth in the vicinity of the transmitter. The results of tests in which the angle of inclination of the antenna was varied, while the antenna remained in the plane containing the transmitter and receiver, are given in Fig. 12.

The theoretical radiation characteristics of antennas in "principal" positions, i.e., horizontal or vertical, over perfectly

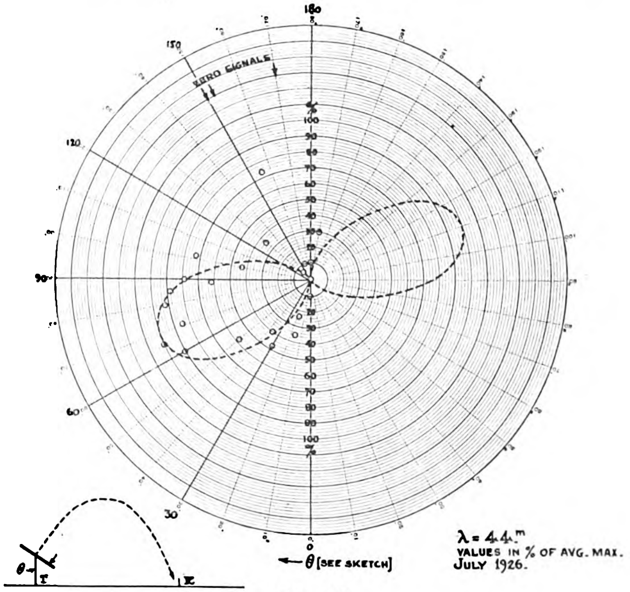

Fig. 12

conducting earth, have been considered by various investigators.[10] The problem of the antenna over an imperfect earth has been attacked also.[11] The problem of an inclined antenna offers many difficulties, even if perfect earth conditions are assumed. In this

[10] G. W. Pierce, "Electric Waves and Oscillations." S. Ballantine, "On the Radiation Resistance of a Simple Vertical Antenna at Wavelengths below the Fundamental," Proc. I. R. E., **12**, 823; December, 1924. S. Ballantine, "On the Optimum Transmitting Wavelength for a Vertical Antenna over Perfect Earth," Proc. I. R. E., **12**, 833; December, 1924. S. A. Levin and C. J. Young, "Field Distribution and Radiation Resistance of a Straight Vertical Unloaded Antenna Operating at One of its Harmonics," Proc. I. R. E., **14**, 675; October, 1926.

[11] L. Bouthillon, "Influence de la nature du sol sur l'emission et la reception radio-electriques," *L'Onde Electrique*, **6**, No. 71; November, 1927.

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case the earth surface was composed of fine sand, which was highly piezo-electric,[12] showing strong resonances at frequencies from as high as 15,000 kc to less than 3,000 kc, that is, over a range of frequency considerably greater than that employed in the experimental work. One is then confronted with the vexatious though interesting problem of trying to account for the effect of the dancing sands under the atenna in attempting to calculate the radiation characteristics.

When using this antenna system in long range communication, it was found that the position of the antenna had no marked effect, though no careful measurements could be carried out. Apparently the random polarization of the waves at the receiver, due to the characteristics of the medium of propagation, completely wiped out all effects due to the position of the antenna itself.[13] These longer range tests were carried out mainly with amateurs in England and Belgium.

In conclusion the author wishes to express his appreciation of the able assistance and active cooperation of Mr. Walter D. Siddall and Mr. Gordon G. Macintosh in carrying out the experiments.

[12] I am indebted to Dr. G. W. Pierce for the suggestion that the piezo-electric character of the sand might be an important influence.

[13] Baker and Rice, loc. cit.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

WIRELESS TELEGRAPHY AND MAGNETIC STORMS*

By
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Summary—*A recent theory of auroras and magnetic storms attributes these phenomena to the action of a flash of ultraviolet light from the sun. The flash causes an unusual ionization in the Kennelly-Heaviside layer. Therefore, it is only daylight wireless circuits which are, or may be, disturbed at the commencement of the magnetic storm, the night circuits remaining normal until dawn when they may be disturbed; the disturbance in the daytime circuits may persist after night-fall. This very simple theory is found to be borne out in a detailed discussion of the data of the short-wave (15 to 40 meters) circuits of the United States Navy during the magnetic storms of May 28, July 7, October 18, and October 24, 1928.*

IT is an accepted belief, based on a long accumulation of evidence, that terrestrial magnetic storms are caused by unusual bursts of radiation of some sort from a disturbed region of the sun. The assumption that the solar disturbance is a flash of ultraviolet light has been shown in a recent paper[1] to explain practically all the complicated magnetic effects observed during a magnetic storm. In a simple case the flash was assumed to blaze out at full intensity for some minutes and then to die away in a few hours or days; it might, of course, flare out again irregularly or intermittently. The flash was such as would come from a hot spot 1/10,000 of the solar disk in size and at a temperature of 30,000 degrees K; the radiation from the hot spot would be largely in the far ultraviolet region of wavelengths which ionize the atmospheric gases. Calculations showed that this radiation, barely perceptible at sea level because it is almost entirely absorbed in the upper reaches of the atmosphere, would heat up the atmosphere at levels above 50 km causing it to expand, and would increase the ionization in the high atmosphere very greatly. The effect of the flash was therefore to increase the ion and electron densities in the Kennelly-Heaviside layer and to raise the entire layer upward about 100 km. This agreed well enough with the measure-

* Dewey decimal classification: R113.5. Original manuscript received by the Institute, December 28, 1928. Published by the permission of the Navy Department, Washington, D.C.

[1] H. B. Maris and E. O. Hulburt, "A Theory of Auroras and Magnetic Storms," *Phys. Rev.*, March, 1929.

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ments of Dahl and Gebhardt[2] who found an increase in the apparent height reached by 70-m waves in the daytime of about 70 km during the magnetic storm of August 19, 1927, and with the recent measurements of Hafstad and Tuve,[3] who observed an increase in the apparent daytime height of the layer of 100 km during the magnetic storms of October 7 and of October 18, 1928.

The atmosphere on the daylight side of the earth is influenced directly by the flash, that on the night side is not affected directly but indirectly, as the day side is turned by the rotation of the earth and carries with it into the shadow the influence it received while in the sunshine. Therefore, only daytime wireless communication may be disturbed during the commencement of the magnetic storm, the night communication remaining normal. If the flash continues, with perhaps lessening intensity, the night communication will, or may, experience disturbance at dawn. If the communication channel extends from night into day, or vice versa, the effects may be complicated. Since the flash heats the atmosphere and thereby raises the Kennelly-Heaviside layer in the sunlit areas, long-wave daylight transmission may be improved, or probably will not be weakened, during the magnetic storm, because the waves are propagated with fewer earth and layer reflections than normally. At night, however, due to the diffusion of the unusual amount of ionization into lower strata where the gas molecules are more dense, the absorption of the long waves is increased and the transmission falls below the normal night-time values. These conclusions are in general accord with the facts brought out by Austen,[4] Espenschied, Anderson and Bailey,[5] Pickard,[6] Anderson,[7] and others for waves from 5000 to 20,000 m.

In general, long waves are not so sensitive as the short waves to movements and changes in the upper atmosphere, because of their length and their relatively less penetration into the ionized regions. Therefore, the very superficial description just given of the effects of the storm in the high atmosphere,

[2] O. Dahl and L. A. Gebhardt, Proc. I. R. E., **16**, 290; March, 1928.
[3] Hafstad and Tuve, *Terr. Mag. and Atmos. Elec.*, **34**, March, 1929.
[4] L. W. Austin, Proc. I. R. E., **15**, 825; October, 1927.
[5] L. Espenschied, C. N. Anderson, and A. Bailey, Proc. I. R. E., **14**, 7; January, 1926.
[6] G. W. Pickard, Proc. I. R. E., **15**, 749; September, 1927.
[7] C. N. Anderson, Proc. I. R. E., **16**, 297; March, 1928.

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which has perhaps sufficed for a rough explanation of the behavior of the long waves during the storm, is hardly sufficient for the case of the short waves. A more careful examination is necessary, especially as data are now available in considerable detail of the short-wave storm characteristics. The conditions in the atmosphere arising from the solar ultraviolet flash are in the main complicated. If, as a result of the heavy ionization caused by the storm, the lower surface of the ionized layer is sharply defined, the short waves may be reflected without great penetration and their transmission will be good; but if the lower surface is not well marked, or if it is agitated, etc., the transmission may be greatly disturbed. And which of the conditions obtains in any particular case depends upon the rates of diffusion of the ions and electrons, the winds in the high atmosphere, the character of the solar flash, e.g., whether it is short and intense, or mild and long, or irregular, etc., etc. Actually the various possibilities are met with, as brought out in the detailed discussion of four magnetic storms given in the following paragraphs, there being cases in which the short-wave full daylight transmission was greatly disturbed at the commencement of the storm, and other cases in which it was disturbed to a less extent. But, it must be emphasized again, no case was encountered in which the full night-time communication was troubled perceptibly at the commencement of the storm. Since the atmosphere is heated and expanded at the beginning of the storm in those regions where the sun is more or less overhead, as from ten to four o'clock local solar time, winds in the high atmosphere will blow from the heated regions and produce turbulence and swirls of the ions and electrons on the borders of those regions. Short wireless waves passing into this turbulence will be scattered, or absorbed, or poorly reflected, as the case may be, and poor transmission will result. When night falls over the daylight areas, the turbulence may slowly diminish and the transmission may improve in the early morning hours, and then with the dawn may continue to improve or may fall off again, depending upon the activity of the flash. Thus, in part, the solar disturbance, which causes the magnetic storm, accentuates and widens the usual sunset and sunrise vagaries in the wireless transmission.

In Figs. 1, 2, 3, and 4 are given data for the four magnetic storms of May 28, July 7, October 18, and October 24, 1928.

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respectively. The wireless information was obtained from the log books of the Radio Division of this Laboratory, and the magnetic data from the Coast and Geodetic Survey Observatory at Cheltenham, Maryland, U. S. A. The abscissas of the figures are the Eastern Standard times of the various days. The ordinates give the wireless communication as expressed qualitatively by the words "good", "poor", and "bad". "Good" means the normal signal intensity obtained with the usual day and night routine wavelength bands which were 14 to 25 and 25 to 40 meters, respectively; "poor" means signals below normal due to fading and low intensity, and "bad" means signals very much below normal, or absent altogether. Curves 1 of the figures refer to the short-wave communication from Washington, D. C., U.S.A., to Europe, along the Atlantic sea-board and to South America. Curves 2 refer to the Washington-San Francisco short-wave circuit.

Curves 3 of the figures give the variations in *H*, the horizontal component of the magnetic field of the earth; these curves are free-hand sketches from the magnetograph records of the Cheltenham Observatory. On a quiet day the *H* curves are smooth with diurnal undulations of a regular type. A ragged *H* curve with a sharp initial rise, a fairly rapid decrease to negative values and a slow period of recovery, is the characteristic type of storm curve. These characteristics are world-wide, being the same, apart from local variations peculiar to the station, for all the magnetic stations of the earth; the storm commences simultaneously, within a minute at all the stations. According to the theory of auroras and magnetic storms[1] the initial rise and fall to zero of the *H* curve is due to a pulse of current flowing from west to east around the earth in the conducting region of the atmosphere. The pulse of current is caused by the increased ionization in the Kennelly-Heaviside layer when the ultraviolet flash flares up to full intensity. The decrease of *H* to negative values is attributed to the induced currents in the earth, and the slow recovery of *H* to the dying away of the currents as well as to the fading out of the flash. Irregularities which occur during the recovery period probably indicate recurrences of the flash, although the induced current reactions may also play a part.

The four storms illustrate four quite different types of conditions; the effects on wireless communication, however, were

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in each case in general accord with the theoretical views which have been outlined. On May 27 the magnetic disturbance began gently at about 10 A.M., as shown in Fig. 1, increased to a moderate intensity on May 28 and died away the next day.

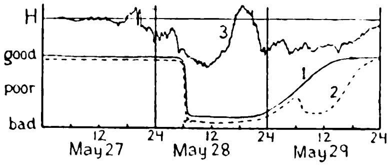

Fig. 1

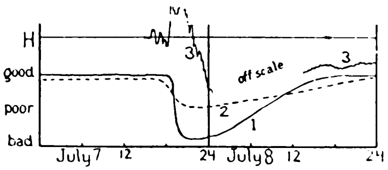

Fig. 2

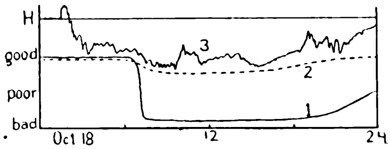

Fig. 3

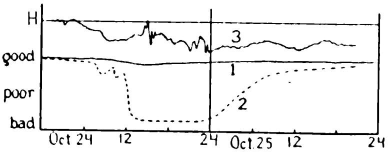

Fig. 4

In the above figures the abscissas are Eastern Standard Time. Curve 1 gives the short-wave communication on the Washington circuits to Europe and South America and along the Atlantic seaboard; curve 2 refers to the Washington-San Francisco circuit. "Good" means the normal signal strength. Curve 3 gives *H*, the variations in the horizontal component of the earth's magnetic field, as sketched from the magnetograph records of the Cheltenham Observatory, Maryland, U.S.A.

Wireless showed no evidences of trouble on the 27th, but at about 6 A.M. of the 28th the wireless channels both to the east and the west became inoperative. During the day of May 28 wireless was poor, and then began to recover in the night. The

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recovery on the Atlantic circuits continued through the dawn of May 29 and was complete by noon. The San Francisco communication, however, became poor again at sunrise of May 29 and did not return to normal until that night. Just why the San Francisco circuit experienced more trouble on May 29 than the Atlantic circuits is not clear. Perhaps local winds in high atmosphere, irregularities in the ionization, or many other factors neglected in the present simple theory were the cause. It should also be mentioned that the wireless data indicated in a number of cases a non-reciprocity of signalling, that is, signals were received at one station from a second, but the second station did not receive the signals from the first. However, the data were not sufficiently complete, nor is the theory sufficiently detailed, to permit the discussion of these complexities. They will require further investigation.

On July 7 at 4:30 P.M. a short severe magnetic storm set in and continued until about noon on July 8, as shown in Fig. 2. Communication from Washington both east and west dropped sharply to low values at about 7 P.M., and recovered slowly during the night and the following day. The Pacific circuits Honolulu to Cavite, which lay nearly directly under the influence of the storm flash, apparently were not disturbed on July 7 (actually July 8 at Cavite, which is on the other side of the international date line), but at 8 A.M., Honolulu local solar time, of July 8 short-wave communication faded to low values for ten hours.

The storm of October 18, Fig. 3, began at 2:27 A.M., with a fairly pronounced magnetic disturbance; the Washington circuits to the east and west showed no indications until the dawn. During the daylight hours of the 18th the Atlantic communication was poor or zero; the San Francisco circuit, although somewhat troubled, remained in operation.

A storm of moderate intensity, Fig. 4, began slowly in the morning hours of October 24 with a small increase at about 2 P.M. The Washington-San Francisco circuit had periods of low signal strength from 9 A.M. to 12, and in the afternoon communication became poor. The Atlantic circuits were little disturbed. Meanwhile the Pacific circuits, which were in the dark when the storm began on the 24th, experienced low signal strength and fading during the daylight hours of this period.

In addition to the facts which have been mentioned, the

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wireless log books indicated clearly that fading and poor signal strength were the general rule on days of magnetic character 1 or greater; 1 means moderate, 2 strong magnetic disturbance, etc. In 1927, a year of maximum disturbance, there were 83 days of character 1 or greater; wireless therefore might have encountered difficulties due to solar activity on more than one fifth of the days of the year. Three or four years from now the solar activity will be at the minimum of its eleven-year cycle and one may expect a less number, perhaps only 10 or 20, of magnetically disturbed days during the year. The magnetic observations may properly be regarded as the weather stations for the wireless traffic and the magnetic storm charts as the wireless weather maps.

In conclusion it is a pleasure to express our thanks to the Coast and Geodetic survey for their courtesy in giving us promptly and fully the magnetic data of their observations.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

RECENT DEVELOPMENTS IN SUPERHETERODYNE RECEIVERS*

By

G. L. Beers[1] and W. L. Carlson[2]

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Summary—*Major electrical elements of a modern superheterodyne receiver-tuned radio-frequency amplifier, intermediate frequency amplifier detector and audio-frequency amplifier—are briefly discussed in light of recent developments. A practical automatic volume control is described. Curves illustrating the major performance characteristics of the receiver are shown.*

Introduction

In the past four years broadcasting conditions have changed so greatly that receivers which were giving satisfactory service at the beginning of that period are now obsolete. The steady increase in the number of broadcasting stations and the advent of the super-power stations have imposed exacting selectivity requirements on modern receiving sets. The congested condition of our cities with their numerous apartment buildings and the resultant lack of antenna facilities have likewise created the demand for receivers having sufficient sensitivity to permit the use of either a small indoor or outdoor antenna. The marked improvement in the quality of radio programs and their transmission and the development of the moving coil type of loudspeaker have made the fidelity of present day receivers a consideration of major importance. The superheterodyne receiver is particularly adapted to meet these modern broadcasting conditions. It is the purpose of this paper to discuss some recent developments in this type of receiver.

General

The conventional tuned radio-frequency receiver has two main electrical elements, namely, the tuned radio-frequency amplifier and the detector and audio-frequency amplifier. The modern superheterodyne receiver has both these elements and, in addition, an intermediate frequency amplifier with associated

*Dewey decimal classification: R343. Original manuscript recieved by the Institute, November 14, 1928. Presented before New York meeting of the Institute, March 6, 1929.

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frequency converting oscillator and detector. The high amplification and great selectivity of this intermediate frequency amplifier largely account for the remarkable performance of the superheterodyne receiver.

The ease of obtaining high amplification and selectivity in an intermediate frequency amplifier is chiefly due to the low frequency used and to the fact that the characteristics of such an amplifier are independent of the broadcast frequency to which the set is tuned. At these lower frequencies by the use of coupled circuits it is possible to obtain the so-called band-pass filter selectivity characteristics.

Improved radio-frequency circuits and the use of a higher intermediate frequency have contributed a considerable re-

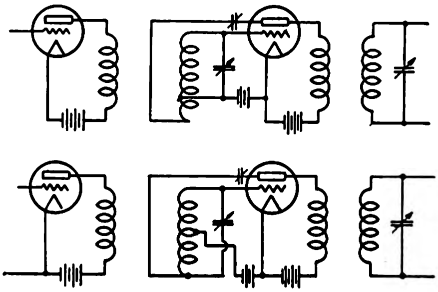

Fig. 1

duction in hiss and extra responses common to previous superheterodyne receivers.

The adjacent channel selectivity and the fidelity of broadcast receivers using a given number of tuned circuits are so related that it is impossible to increase either of these characteristics beyond certain limits without a corresponding sacrifice in the other.

A radio receiving set may be designed using four tuned circuits which will have a better adjacent channel selectivity than an eight tuned circuit receiver which is designed for good fidelity. The four tuned circuit set will be lacking in fidelity due to the attenuation of the side bands, and will also have less selectivity for stations separated by 30 kc or more. If the four tuned circuit set is made to equal the fidelity of the eight tuned circuit set, it will be inferior in selectivity in every way.

The use of broader radio-frequency and intermediate frequency circuits in connection with a negatively biased detector

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capable of high voltage outputs and a single stage of audio-frequency amplification has contributed a major improvement in the fidelity of the modern superheterodyne receiver.

The incorporation of a practical automatic volume control has eliminated the necessity of repeated adjustment of the volume control both when tuning from distant to local stations or vica versa, and when receiving stations whose field strength is varying periodically.

Radio-Frequency Amplifier

The selectivity of a well designed superheterodyne receiver for eliminating local stations is chiefly determined by the selec-

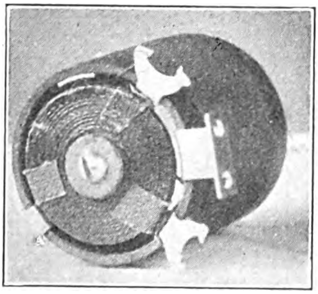

Fig. 2

tivity characteristics of the intermediate frequency amplifier. There are, however, a few exceptions; for example, if no radio-frequency selectivity is employed, two signals of equal strength differing in frequency by twice the intermediate frequency will be received with equal intensity with the heterodyne oscillator tuned midway between them. In addition to this objectionable response, harmonics of the oscillator heterodyning undesired signals are likely to cause interference. An input circuit tuned to the signal it is desired to receive reduces the possibility of interference to some extent, but in the vicinity of powerful broadcasting stations more than one tuned circuit is necessary.

The usual type of tuned radio-frequency transformer consisting of a few turns in the plate circuit of the amplifier tube coupled to the tuned secondary gives a selectivity characteristic which is much sharper at the low-frequency end of the broadcast range than at the high. In fact, it is usually so sharp that two

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or three stages of this type of amplification results in considerable reduction of the high modulation frequencies. A transformer of this type designed for good selectivity at the high-frequency end of the range will usually be lacking in amplification at the low-frequency end of the range.

A new radio-frequency system has been developed which overcomes most of the objections to the type just described. It provides much more uniform selectivity and amplification without resorting to some mechanical means of varying the coupling with frequency. The only difference between the new transformer and the old is in the type of primary used. Where the primary for the old type of radio-frequency transformer was

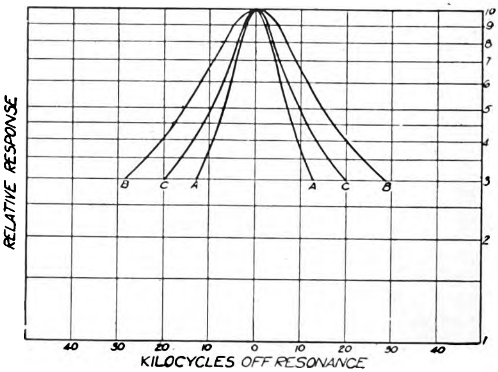

Fig. 3

wound with a small number of turns and was resonant to a frequency above the high-frequency end of the broadcast range, the improved transformer has a large number of turns on the primary, making it resonant to a frequency below the low-frequency end of the range. The old primary increased the effective secondary resistance at the higher frequencies where it was normally too great for good selectivity. The new primary increases the effective secondary resistance at the low-frequency end of the range where it is normally too low for high fidelity.

Since the primary is resonant at a frequency below the broadcast range, the amplification is increased at the low frequencies and reduced somewhat at the high, making it uniform over the range.

In order to realize the normal amplification of high inductance primary transformers, some means must be used to compensate

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for the effect of the grid to plate capacity. This is due to the primary being tuned to a lower frequency than the secondary, thus giving the plate circuit capacitive reactance. The voltage fed back through the tube capacity is therefore of such a phase as to oppose the applied grid voltage and will reduce this voltage to a fraction of its normal value. Either of the methods shown in Fig. 1 may be used to overcome this effect. If the feed-back capacity is made too large, it is possible to make the circuit oscillate.

Fig. 2 shows one form of a transformer of the high inductance primary type. Curves *A*, *B*, and *C*, Fig. 3, are resonance curves taken at 600 kc showing the effect of the two types of primaries

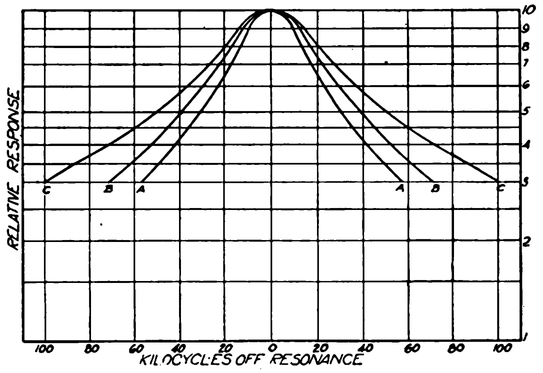

Fig. 4

on the tuned secondary. Curve *A* is the resonance curve of the secondary alone. Curve *B* shows this secondary coupled to a high inductance primary, and Curve *C* shows the same secondary coupled to a low inductance primary. Curves *A*, *B*, and *C*, Fig. 4, are similar curves taken at 1400 kc. Fig. 5 shows the amplification curve for a complete stage of radio-frequency amplification using a high inductance primary transformer and a UY-227 tube.

Intermediate-Frequency Amplifier

In past superheterodyne receivers the intermediate frequency has usually been in the neighborhood of 40 or 50 kc. This choice resulted from the ease of obtaining a stable amplifier for a frequency in this region having the necessary amplification and the desired selectivity characteristics. Now, however,

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another important factor must be considered in the choice of an intermediate frequency.

From Fig. 6, showing the selectivity of the two tuned radio-frequency circuits, it is noted that the higher the intermediate

Fig. 5

frequency used the less the possibility of encountering interference from stations separated by twice the intermediate frequency. From the curve it will be seen that this interference for a 40-kc amplifier would be 350 times that for a 400-kc amplifier and 60 times that for a 200-kc amplifier.

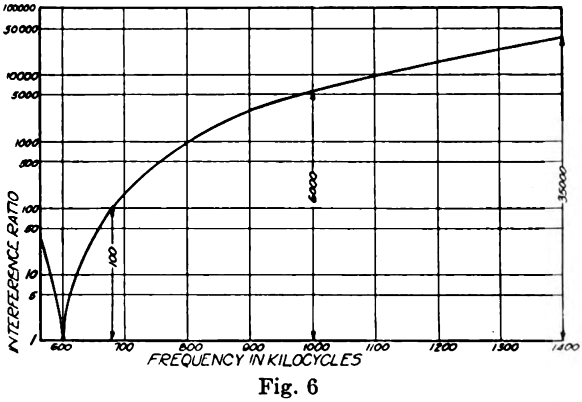

Fig. 6

An intermediate frequency of 180 kc was finally chosen as the best compromise between amplification, stability, selectivity, and undesired responses.

With an intermediate frequency of 180 kc and the oscillator tuned to a higher frequency than the radio-frequency circuits,

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it will be seen that broadcasting stations separated by twice the intermediate frequency will not produce interference at frequencies above 1140 kc. The selectivity of the radio-frequency system at the high frequencies is therefore not as important in this receiver as at the low frequencies.

Both the primary and the secondary of the three transformers used in the 180-kc amplifier are tuned, and the two windings are so coupled as to give a broad top resonance characteristic. Curve *A*, Fig. 7, is for a single stage of the inter-

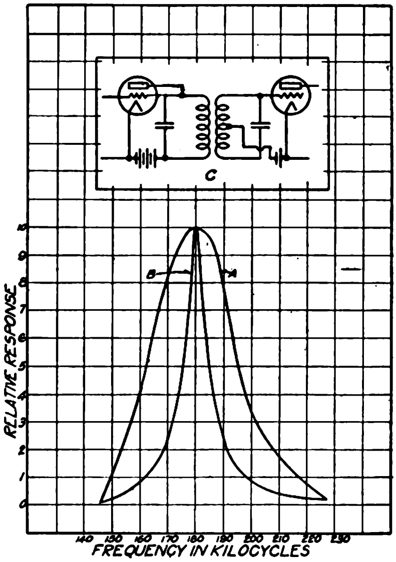

Fig. 7

mediate-frequency amplifier as shown in circuit *C*, having the proper magnetic coupling between the primary and secondary Curve *B* shows the combined characteristic of the primary and secondary used as separate tuned circuits with vacuum-tube coupling between them. Such a characteristic would be obtained if these individual circuits were used separately in a cascade amplifier. These two curves indicate the improvement that can be obtained by the use of coupled circuits, provided the losses in the individual circuits can be kept reasonably low.

The overall resonance curve for the intermediate-frequency amplifier is shown in Fig. 8. The approximately ideal band-pass filter characteristics of this amplifier should be noted, the band

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width at 50 per cent peak amplitude being 16 kc while at 1 per cent peak amplitude it is only 40 kc.

The three transformers are each mounted in individual metal containers which serve both to protect the transformer and shield it electrically. The primary is tuned by a compact fixed condenser, and the secondary by a small adjustable condenser which permits accurate alignment of the three transformers.

The balancing, primary, and secondary tuning condensers are mounted on a small piece of bakelite to which are also riveted

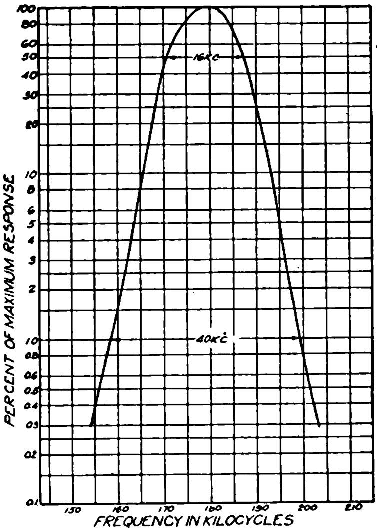

Fig. 8

the two brackets for supporting the transformer windings. Fig. 9 illustrates the manner in which these condensers, the transformer windings, and the terminals are mounted. A transformer completely assembled ready to mount on the chassis is shown in Fig. 10.

Audio-Frequency System

The audio-frequency detector in radio receiving sets is responsible for some distortion and contributes to such disturbances as microphonic howl and a.c. hum.

These objectionable features are considerably reduced, and other performance advantages are gained by employing a plate circuit detector and by using but one audio stage.

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The relative merits of the conventional audio system consisting of a grid circuit detector and two audio stages as compared with a combined radio and audio system employing a radio-

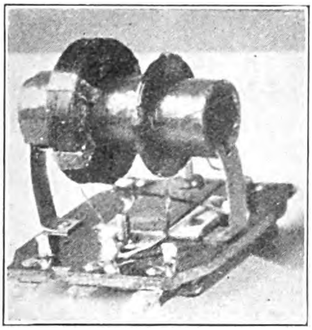

Fig. 9

frequency stage, plate circuit detector, and one audio stage are as follows:

Changing from a grid circuit to a plate circuit detector in a given receiver results in a sacrifice in sensitivity. This reduction in sensitivity can be overcome by increasing the radio-frequency

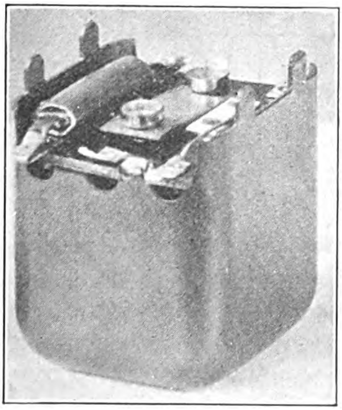

Fig. 10

amplification and reducing the audio-frequency amplification. In general, it can be said that substituting a radio-frequency stage of equal amplification for an audio stage will result in no loss of sensitivity from substituting a negatively biased detector for a grid leak and condenser detector.

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With improved audio response, particularly at low frequencies, when two audio stages are used, it has become increasingly difficult to avoid a.c. hum from the power supply, and audio howls due to the microphonic action of tubes. The blasting and breaking up of the sound output when tuning through a local station is ordinarily due to overloading the output tube. This disturbance can be reduced if the circuits are so designed as to allow the detector and output tube to overload at the same time.

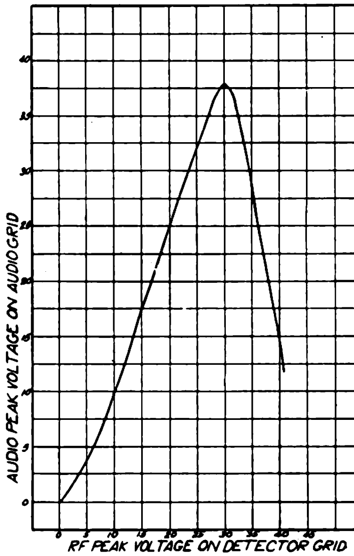

Fig. 11

The time lag distortion associated with detectors employing grid leak and condenser combinations is not present when plate circuit detection is used.

The curve in Fig. 11 shows the relation between the peak voltage on the grid of a UY-227 tube functioning as a negatively biased detector and the peak audio voltage on the grid of the audio amplifier tube. The plate potential on the UY-227 tube is 180 volts and the bias 25 volts. The carrier is modulated 15 per cent at 100 cycles. The reduction in output for an input in excess of 30 volts is due to the grid current load on the tuned circuit, through which the voltage is applied to the detector grid.

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AUTOMATIC VOLUME CONTROL

A receiving set equipped with a practical automatic volume control has several distinct advantages over the set which lacks this equipment.

One of the chief advantages is the ability to change the tuning of the receiver without making any change in the volume control setting. Once the volume control has been set for the desired sound output, either distant or local stations may be tuned in without any further volume control adjustment. If the distant station provides sufficient field strength (depending

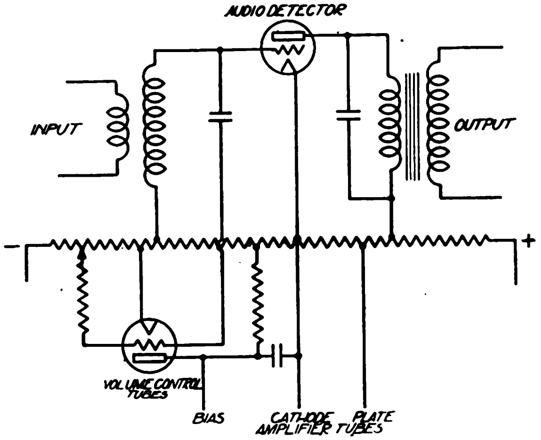

Fig. 12

on the maximum sensitivity of the receiver) it will produce practically the same sound output as the local station. It is assumed that the per cent modulation used by both stations is nearly the same.

The ability of an automatic volume control to limit the sound output to any desired value is of particular importance in light of the present trend towards the use of output tubes capable of handling considerable power.

While automatic volume controls have sometimes been called fading eliminators, this term is, of course, erroneous. An effective automatic volume control, however, does automatically adjust the sensitivity of the receiver while the field strength from a station is varying so as to maintain a constant output

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There are numerous occasions when a powerful distant station has sufficient field strength at any time during an evening to give a satisfactory loudspeaker signal, but the field strength varies through such a wide range and so frequently that the program cannot be received satisfactorily without continuous adjustment of the volume control. Under such conditions an automatic volume control will function satisfactorily. The only indication that the user will have that the signal is fading will be an increase in the ground noise when the signal drops to such

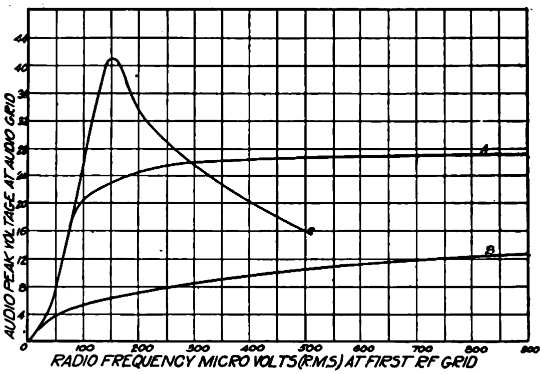

Fig. 13

a value that nearly the maximum sensitivity of the receiver is required to produce the desired sound output.

There are some types of fading that are accompanied by a distortion of the audio modulation, and in such cases an automatic volume control will be of little value as far as compensation for this type of fading is concerned.

The chief objections to past automatic volume control systems have been the number of adjustments required and the use of separate voltage supplies for certain parts of the circuit. To obtain sufficient control some systems also require additional amplification either for the a.c. voltage on the detector grid or for the bias voltage before it is impressed on the amplifier grids. This latter objection is overcome by the use of but one stage of audio amplification and the corresponding increase in the grid swing on the detector tube. Fig. 12 shows a schematic diagram for an arrangement which overcomes the other objections. The grid of the volume control tube is connected in parallel through a

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coupling condenser to the grid of the second detector. The voltage drop across a resistor in the plate circuit of this tube is used as additional negative bias on the amplifier tubes and thus reduces the sensitivity of the receiver. The values of this plate circuit resistor and its shunting capacity are so chosen as to give the combination a time constant sufficiently low to prevent the smoothing out of the lower modulation frequencies and still high enough to prevent its action from being sluggish.

The bias on the volume control tube is normally adjusted to the point where, when no signal is being received, the tube

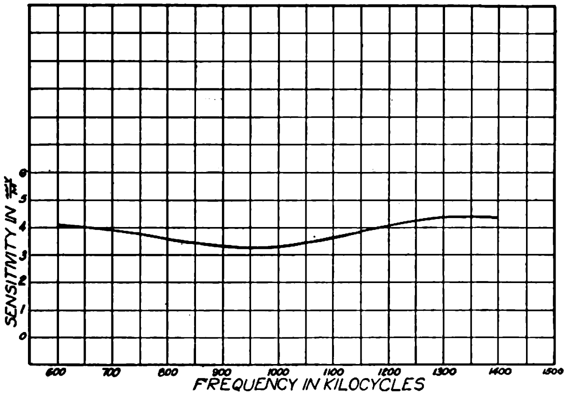

Fig. 14

draws no current. Then when the grid swing on the second detector exceeds a certain value, the effective bias on the control tube is reduced and its plate current increased. This increase in plate current causes a corresponding increase in the bias voltage applied to the radio-frequency and intermediate frequency amplifier tubes with a resultant reduction in the sensitivity of the receiver, thus tending to maintain a uniform grid swing on the second detector. By means of a manual control the negative bias on the volume control tube can be increased permitting a larger grid swing on the detector tube before the automatic volume control tube takes effect. In this manner any desired audio output can be obtained. The variation of the bias on the volume control tube as the manual adjustment for the output level also compensates for the manufacturing variations in cutoff of the tubes as automatic volume control tubes. Curves *A* and *B* in Fig. 13 show the effectiveness of the automatic

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volume control for two output levels. Curve *C* shows the variation in output without the automatic volume control.

The use of an automatic volume control in a radio receiver presents some problems which are not encountered when this equipment is not used. Due to the automatic volume control maintaining constant audio output, it is difficult to tune the receiver accurately unless some resonance indicating device is provided. A meter in the plate circuit of either the control tube or the tubes controlled provides a satisfactory resonance indicator.

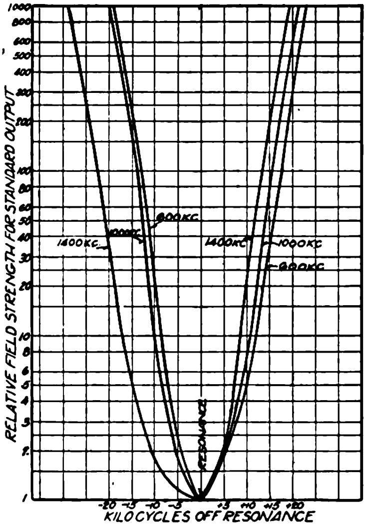

Fig. 15

In localities where there is considerable interference due to power line leakage, etc., the noise encountered when tuning between stations with a set equipped with automatic volume control is objectionable unless a sensitivity control is provided. This is due to the fact that when no signal is being received the receiver is automatically adjusted for maximum sensitivity, and the time constant of the automatic volume control circuit is such that it does not limit the sharp impulses of such interference. The sensitivity control may be a voltage divider which varies the normal bias on the amplifier tubes, or in receivers where an untuned input circuit is employed a voltage divider may be used to vary the signal potential on the grid of the input tube.

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By means of the sensitivity control, the maximum sensitivity of the receiver can be reduced to such a value that the noise encountered in tuning between stations is not objectionable. This sensitivity control will in no way destroy the effectiveness of the automatic volume control.

Overall Performance

The major electrical elements of a modern superheterodyne receiver have been briefly discussed. The performance char-

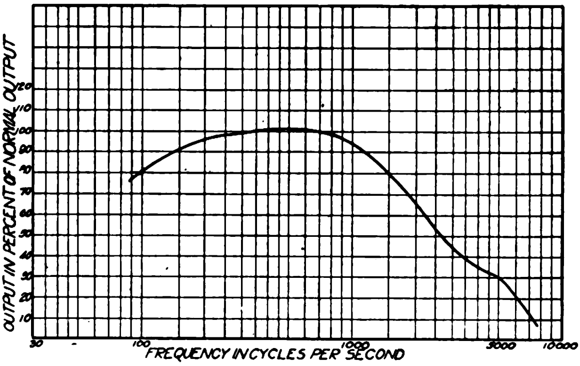

Fig. 16

acteristics of a complete receiver embodying these elements are as follows:

Fig. 14 is the sensitivity curve of a receiver consisting of two radio-frequency stages, heterodyne detector and oscillator, two intermediate-frequency stages, high output detector, and single audio stage. The receiver employs UY-227 tubes throughout except for the power output tube.

The three curves in Fig. 15 show the selectivity of the receiver at three frequencies in the broadcast range. The fidelity characteristic of the complete receiver is shown in Fig. 16.

The curves show the high uniform degree of selectivity obtained by the use of the eight tuned circuits while still retaining unusual fidelity characteristics.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

AN EXTENSION OF THE METHOD OF MEASURING INDUCTANCES AND CAPACITIES*

BY
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Summary—*The substitution method commonly employed for measuring small capacities is shown to be a special case of a more general principle. As other special cases of this principle methods are presented for simultaneously measuring inductance and capacity when joined in series and when joined in parallel. The cases discussed indicate the method of application of the general principle to any type of measuring or measured circuit.*

THE method commonly employed for measuring the capacity of condensers, in which the unknown condenser is connected in parallel with a calibrated condenser in a resonant circuit, is a special case of the more general method described in this paper. The several special cases of the more general method are of interest in that they furnish simple and convenient ways in which to measure circuit elements when

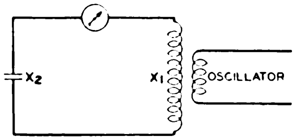

Fig. 1

connected together in circuit. The general principle of the substitution method can be explained by means of Figs. 1 and 2. A series circuit including a current indicator (e.g., a wavemeter) is coupled loosely to an oscillator. Resonance is obtained by adjusting x_2 so that

$$x_2 = x_1 \tag{1}$$

The coupling between x_1 and the oscillator is assumed to be so loose that the reaction between the circuits is negligible. The unknown reactance x_3 is then connected in parallel with x_2, and the latter is then readjusted so that resonance is again obtained. Then

$$\frac{x_3 x_2'}{x_3 + x_2'} = x_1 \tag{2}$$

* Dewey decimal classification: R230. Original manuscript received by the Institute, November 19, 1928.

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Equating (1) and (2), and solving for the unknown x_3:

$$x_3 = \frac{x_2' x_2}{x_2' - x_2} \tag{3}$$

This is the general condition for all such measurements, and from it the particular cases can be derived. The wavemeter is the most convenient instrument to use in many of these measurements, so that we may put

$$x_1 = j\omega L_1, \quad x_2 = -\frac{j}{\omega C_2}, \quad x_2' = -\frac{j}{\omega C_2'} \tag{4}$$

and (3) becomes

$$x_3 = \frac{j}{\omega(C_2' - C_2)} \tag{5}$$

Case 1. If x_3 is a pure capacitive reactance, then equation (5) becomes

$$C_3 = C_2 - C_2' \tag{6}$$

Case 2. If x_3 is a pure inductive reactance, it becomes

$$L_3 = \frac{1}{\omega^2(C_2' - C_2)} \tag{7}$$

Case 3. If x_3 consists of an inductance in series with a capacity, as for instance an antenna circuit, then $x_3 = j(\omega L_3 - 1/\omega C_3)$ and (5) reduces to

$$L_3 = \frac{C_3 + (C_2' - C_2)}{\omega^2 C_3 (C_2' - C_2)} \tag{8a}$$

or,

$$C_3 = \frac{(C_2' - C_2)}{\omega^2 L_3 (C_2' - C_2) - 1} \tag{8b}$$

In order to obtain the values of L_3 and C_3 it is necessary to make two measurements at different frequencies. Let these frequencies be f_1 and f_2, and the corresponding values of $(C_2' - C_2)$ be $\underline{C}'$ and $\underline{C}''$. Then (8a) may be written

$$L_3 = \frac{\underline{C}' + C_3}{\omega_1^2 \underline{C}' C_3} = \frac{\underline{C}'' + C_3}{\omega_2^2 \underline{C}'' C_3}$$

and solving for C_3:

$$C_3 = \frac{(\omega_1^2 - \omega_2^2)\underline{C}'\underline{C}''}{\omega_2^2 \underline{C}'' - \omega_1^2 \underline{C}'} = \frac{1 - \left(\frac{f_2}{f_1}\right)^2}{\left(\frac{f_2}{f_1}\right)^2 \frac{1}{\underline{C}'} - \frac{1}{\underline{C}''}} \tag{9}$$

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Similarly, starting with (8b) and solving for L_3:

$$L_3=\frac{\underline{C}'-\underline{C}''}{\underline{C}'\underline{C}''(\omega_2^2-\omega_1^2)}=\frac{1}{4\pi^2(f_2^2-f_1^2)\left(\dfrac{\underline{C}'\underline{C}''}{\underline{C}'-\underline{C}''}\right)} \tag{10}$$

The quantities $\underline{C}'$ and $\underline{C}''$ may be termed *apparent* capacities. For example, in determining the constants of an antenna, it is regarded as a simple capacity, and its apparent values at two different frequencies are measured as in Case 1. These values,

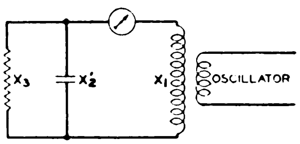

Fig. 2

substituted in (9) and (10), yield the true capacity and inductance. In applying the method to circuits containing coils which have appreciable capacity, the two frequencies should not differ very widely, as the apparent inductance of the coil changes with frequency.

Case 4. If x_3 is an inductance paralleled by a capacity, then

$$x_3=j\,\frac{\omega L_3}{1-\omega^2 L_3 C_3}$$

Following the same procedure as in Case 3, we obtain:

$$C_3=\frac{\omega_1^2\underline{C}'-\omega_2^2\underline{C}''}{\omega_2^2-\omega_1^2}=\frac{\underline{C}'-\left(\dfrac{f_2}{f_1}\right)^2\underline{C}''}{\left(\dfrac{f_2}{f_1}\right)^2-1} \tag{11}$$

$$L_3=\frac{\omega_2^2-\omega_1^2}{\omega_1^2\omega_2^2(\underline{C}'-\underline{C}'')}=\frac{f_2^2-f_1^2}{4\pi^2 f_1^2 f_2^2(\underline{C}'-\underline{C}'')} \tag{12}$$

Discussion

R. R. Batcher[1]: It seems to the writer that several precautions are necessary in applying the measurement methods outlined in this paper. It has been his experience that the procedure taken up in Case 3 and the

[1] Decatur Manufacturing Co., Inc., Brooklyn, N. Y.

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use of formulas (9) and (10) might lead to erroneous results in some cases. This is due to the nature of the circuit resulting when x_3 is replaced by a series arrangement of inductance and capacitance. Such a circuit is resonant to two frequencies, one greater than the natural resonance frequency of the L_3C_3 path and the other less than this frequency. This condition has been examined[2] and used in several cases by the writer and a number of formulas were developed to cover the several variations of conditions found in practice. The following is illustrative of the circuit relations:

$$\frac{f_{(x_1x_2x_3)}}{f_{x_3}} = \sqrt{\frac{L_1C_2 + L_3C_3 + L_1C_3 \pm \sqrt{(L_1C_2 + L_3C_3 + L_1C_3)^2 - 4L_1L_3C_2C_3}}{2L_3C_3}}.$$

This is the ratio of the frequency of $x_3x_1x_2$ paths to the resonant frequency of x_3.

It is the condition that permits this circuit to be tuned to a wavelength shorter than the natural of x_3 that makes the circuit so useful. If x_3 is an antenna system lower waves can be tuned to than can be obtained with a condenser in series with the antenna circuit.

Used in a measurement circuit it is evident that the measurement frequencies must be selected with care, as it will be impossible to obtain a satisfactory adjustment at a frequency near that at which the x_3 path becomes resonant. It can be shown that it is necessary to consider the value of x_1 (that is L_1) in any computations. In some cases it may be desirable to determine the ratio of

$$\frac{\text{Resonant frequency of } x_1x_2x_3}{\text{Resonant frequency of } x_1x_2}$$

instead of that given by the above formula. In this case the values obtained for either the low or the higher frequency must be multiplied by the factor

$$\sqrt{\frac{L_3C_3}{L_1C_2}}$$

which is equivalent to changing the denominator to $2L_1C_2$. It is possible to convert this formula to a more convenient form for use in measuring work.

In Case 4 in the above paper a condition may be found where the resonant frequency of x_3 may be near or equal to the oscillator frequency. In this case there will be little or no change in the tuning of the x_1x_2 circuit after the addition of x_3. It seems that other peculiar conditions might be encountered if the two measuring frequencies happened to be on opposite sides of the frequency of x_3.

Quite aside from the above comments it seems that the accuracy of the above method is determined by the ratio of f_2/f_1. However, mention is made that the two frequencies should not differ very much. If the distributed capacity of the wavemeter pickup coil is measured and the value added to that of the condenser and the precaution taken to use as large a value of C_2 and as small a value of pick up inductance the error is

[2] "Prepared Radio Measurements," pages 89 to 91; and *Wireless Age*, p. 39, April, 1919.

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somewhat lessened and it might be possible to use a larger frequency ratio. Since we are mainly concerned with the difference between two readings of C_2 the effect wavemeter coil capacity which is effectively in parallel with the condenser may not introduce as great an error in results as might be found with the use of two measuring frequencies differing but little. The effect of the distributed capacity of the coil under measurement is rather hard to predict, however.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

APPARENT EQUALITY OF LOUDSPEAKER OUTPUT AT VARIOUS FREQUENCIES*

By
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Summary—*A type of alternation phonometer has been developed which permits rapid switching of power at two frequencies to the same loudspeaker without the distracting effect of the transients that would result from ordinary types of commutation. This result is obtained by the use of rotating condensers to provide variable capacitative reactance in the input circuit of the power amplifier that operates the speaker. The power consumed by the loudspeaker is measured with a specially constructed wattmeter of the electrodynamometer type and the output of the loudspeaker is measured by means of the torques produced on a Rayleigh disc. With the aid of the alternation phonometer and an additional capacitative reactance, the observer is able to adjust the power input to the loudspeaker until two tones of different frequencies appear to have the same intensity. The purpose of the research was to develop a method for the comparison of loudspeaker efficiency at various frequencies that could be used in ordinary laboratories with limited equipment.*

Previous Work

THAT the sensitivity of the human ear is a function of the frequency of the tone has long been assumed, and quantitative results on this sensitivity for *minimum audition* have been obtained by Fletcher and Wegel.[1] They employed a telephone receiver clamped over the ear of the observer as a sound generator. For tones of greater intensity, Dayton C. Miller[2] employed a set of organ pipes which when driven under certain standard conditions of air pressure gave tones which in the opinion of the builder (an expert pipe organ man) gave tones of equal apparent intensity.

B. A. Kingsbury[3] compared tones of different pitch for the sake of determining the actual pressure on the ear drum for equal apparent intensity at various pitches. He employed a single receiver clamped over the ear after the manner of Fletcher and Wegel. The observer varied the intensity of one tone by means

* Dewey decimal classification: R376.3. Original manuscript received by the Institute, November 20, 1928. Presented before meeting of the Buffalo section of the Institute, January 17, 1929.

[1] Fletcher and Wegel, "Frequency Sensitivity of Normal Ears" *Proc.* Nat. Acad. Sci. 8, 5, 1922.

[2] Dayton C. Miller, "The Science of Musical Sounds," Macmillan Co.

[3] B. A. Kingsbury, "A Direct Comparison of the Loudness of Pure Tones," *Phys. Rev.*, **29**, 588; April, 1927.

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of an attenuator, and he shifted his attention from one tone to the other by manually operating a double pole double throw switch so that he listened first to one tone and then to the other.

Previously, Donald MacKenzie[4] had pointed out certain defects in this method of alternating the tones due to the "decay of sensation" on the part of the observer. He therefore devised an instrument called an "alternation phonometer," which bears some relation to the flicker photometer of optics. Power from two oscillators was alternately impressed on a sound generator by means of a polarized telegraph relay driven from another relay at controllable speeds. The speed found to be most satisfactory was approximately 12 per second in the frequency region 100 to 4000. That is, each source was turned on and off 12 times per second. Only about 0.002 second elapsed between the opening of one contact and the closing of the other. A ther-

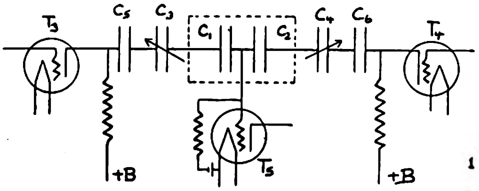

Fig. 1—The Variable Reactance Alternator with Associated Apparatus.

mophone held tightly to the ear was used as the sound generator for part of this work, and an electro-magnetic receiver similarly placed was used for the part involving intensities above the capacity of the thermophone.

Purpose and General Method

The present paper is an account of work done by the authors in attempting to apply the general principle of the alternation phonometer to a method for the determination of the apparent efficiencies of loudspeakers at various frequencies. The general method was to adjust one oscillator to give a steady tone at an arbitrarily chosen intensity. The observer then adjusted the intensity of the other tone until he obtained apparent equality. The two tones were rapidly alternated by a new type of alternation phonometer. The electrical power input to the loudspeaker was then determined for each tone by means of a specially con-

[4] Donald MacKenzie, "The Relative Sensitivity of the Ear," *Phys. Rev.*, 20, No. 4; October, 1922.

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structed electrodynamometer calibrated at each frequency used. The actual sound intensity at the position of the observer's head was determined from Rayleigh disc measurements.

The Alternation Phonometer

In the place of MacKenzie's relay, the authors attempted to use a rotating commutator, but found that with the amplification employed between the commutator and the loudspeaker, the transients introduced by the sudden starting and stopping

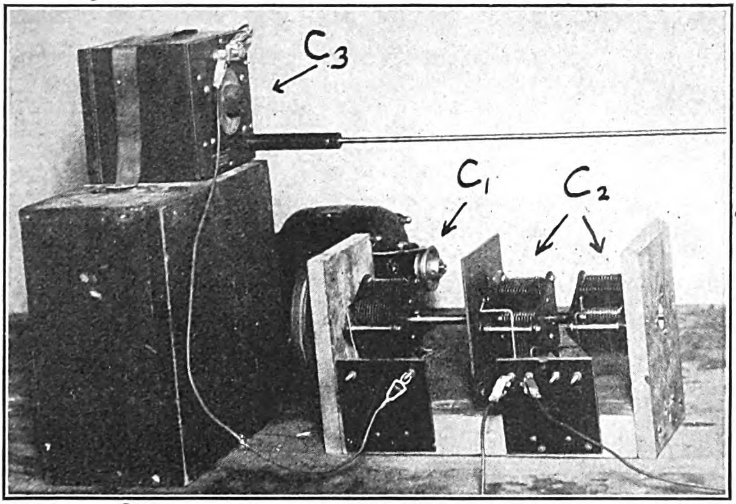

Fig. 2—The Variable Reactance Alternation Phonometer.

of the power from the oscillators were fatiguing to the observer and seriously interfered with his ability to compare the intensities of the tones. An arrangement of variable capacitative impedance in the input of the amplifier was then tried, and it proved to be much more satisfactory.

In the diagram of Fig. 1, C_1 and C_2 are variable air condensers of 0.0005 μf maximum capacity. They are mounted with their rotors on a common shaft so that one condenser is at minimum capacity when the other is at maximum value. Fig. 2 shows two condensers for C_2. These are electrically insulated and are arranged so that they may be used either independently or in series or in parallel. It was found necessary to use only one of them independently in the work reported in this paper. The shaft on which the rotors are mounted has a pulley on one end

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and is belt driven by a variable speed d.c. motor. The speeds most commonly used gave 8 or 9 alternations per second. That is, each tone was heard that number of times per second.

Intensity Control

In the diagram of Fig. 1, C_4 is a variable condenser which enables the operator to make an arbitrary setting for the maximum value of the intensity of the tone to be used as a basis for comparison. C_3 is used to vary the intensity of the tone

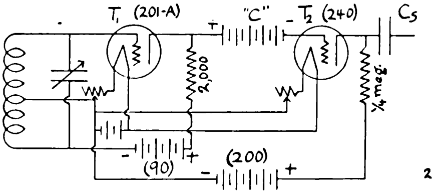

Fig. 3—Oscillator with Conductively Coupled Amplifier.

being compared. It is operated by a long shaft running into the sound compartment so that the observer may readily operate it. C_5 and C_6 were large condensers (4 μf each) used merely to keep the plate potentials off the control condensers C_3 and C_4. T_3 and T_4 were amplifiying tubes which were in turn fed by oscillators.

Oscillators and Amplifiers

The oscillators were of the Hartley type. Air core coils were employed and tuning was accomplished by means of step variable condensers. Each oscillator was conductively coupled to a one-stage amplifier. The arrangement of one of these oscillator-amplifier combinations is shown in Fig. 3.

The controlled output from the rotating condensers C_1 and C_2 was fed into a resistance coupled amplifier of conventional design. The first tube of this amplifier is shown in Fig. 1 and the entire arrangement in Fig. 4. The amplifier is built into a brass-lined box with brass partitions between the separate stages.

Separate filament batteries are used for each oscillator and for the resistance coupled amplifier. But the circuits are so arranged that although separate B and C batteries are used also, the negative sides of all the B batteries used on the amplifiers are electrically connected. This is of course necessary for the operation of the variable reactance in the grid circuit of the amplifier input.

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Loudspeaker

The loudspeaker used in these experiments was an exponential horn driven by a unit of the Baldwin type. The horn was octagonal in cross section and was built of bass wood. The length

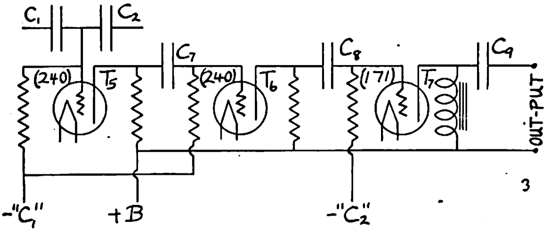

Fig. 4—Resistance Coupled Amplifier. The coupling condensers are 1 μf each. The filaments are in parallel with a common rheostat. Both grid and plate resistances are $\frac{1}{4}$ megohm each.

along the axis was 8 ft. The open end had a cross section of 400 sq. in. and the small end a cross section of approximately 0.3 sq. in. The equation for the shape of the horn is therefore given by

$$A = \epsilon^{(0.075x - 1.181)}$$

where A is the area of cross section in square inches, x is the distance in inches measured along the axis from the smaller end, and ϵ is the base of the natural system of logarithms.

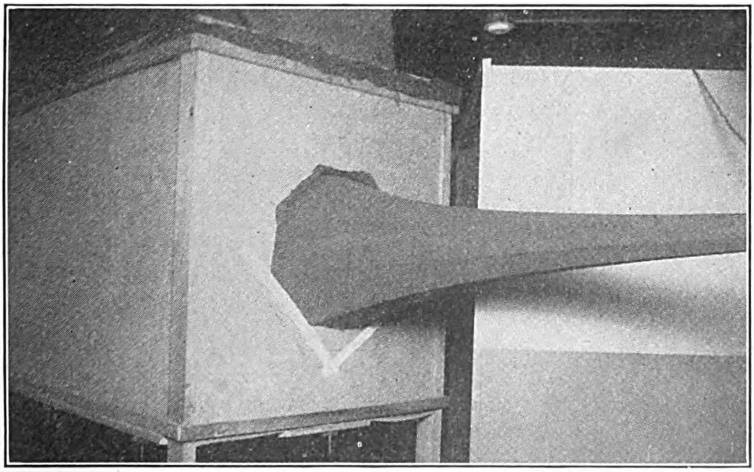

Fig. 5—The Loudspeaker and the Sound Compartment.

Sound Compartment

The mouth of the loudspeaker opened into the end of a 4-ft. x 4-ft.x 8-ft. compartment as shown in Fig. 5. The box was

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built of ½ in. celotex mounted as shown on a 2-in.x 4-in. lumber frame. It was lined on the inside with 2 layers of ½-in. hair felt. On the under side of the box, at the end opposite the loudspeaker, a little of the celotex was cut away, and the hair felt folded back to permit the observer to stand upright on the floor of the laboratory with his head approximately along the mid-axis of the compartment and near its rear end wall. The long shaft from the volume control condenser, C_3, entered the compartment near this end, so that the observer could easily adjust the tone intensity.

Rayleigh Disc Measurements

When Rayleigh disc measurements were to be made the hair felt was closed and the celotex replaced. The Rayleigh disc used in these experiments was a piece of mica 8.75 cm in diameter

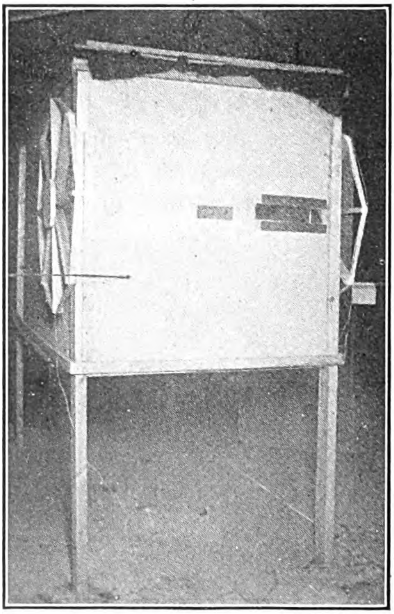

Fig. 6—Rear End of the Sound Compartment Showing the Window for Making Observations and the Octagonal Coils for Establishing the Restoring Field to the Magnet.

suspended from one edge by a thread of unspun silk. On the mica was mounted a small galvanometer mirror which received and reflected light through a narrow glass window in the rear end of the box. (See Fig. 6.) Since the zero position for the disc

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was 45 deg. to the axis of the box, it will be seen that this window was not directly behind the disc, but rather near one edge of the end wall.

The disc also carried a tiny magnet (1.49 cm in length and 0.1389 gram weight) mounted at 45 deg. to the plane of the disc. With the major axis of the sound compartment running north and south, the disc was then in its zero position (45 deg. to the axis of the compartment) when the small magnet was parallel to the earth's magnetic field (H_e).

Essentially a Rayleigh disc depends for its operation on the fact that when an object finds itself in a field of uniform flow it tends to set itself with its greatest dimension cross-wise to the direction of flow. A quantitative development[5] for the case of a round thin disc located in a field of alternating flow where the alternations are simple harmonic gives

$$\text{Torque} = 8/3\pi^2 a^2 n^2 \rho C^3 \sin 2\theta$$

where

a is the amplitude of the vibration
n the frequency
ρ the density of the medium
C the radius of the disc
θ the angle between the normal to the disc and the direction of flow.

Now since the mean square velocity in simple harmonic motion is given by $2\pi^2 a^2 n^2$, and since the energy in a sound wave is proportional to the mean square velocity of the vibrating particles it follows that the torque on the Rayleigh disc is proportional to the intensity of the sound at the point considered. Obviously this torque is a maximum when the angle θ equals 45 deg.

The sound waves striking the disc tended to rotate it and the attached magnet. As the magnet moved out of the earth's magnetic field, the magnetic couple due to that field tended to restrict the motion. A second magnetic field, H, at right angles to the earth's field was then established by sending a measured current through the coils shown in Fig. 6. The couple due to this field also tended to return the magnet (and hence the disc) to its original position. Since the couple due to the earth's field is equal to zero when the magnet is in its original position the final magnetic couple, G, is given by

[5] Winklemann, *Handbuch der Physik.*, II. p. 238.

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$$G = H\ M$$

where H is the field due to the current in the coils and M is the magnetic moment of the magnet. But this couple must just balance the torque on the disc due to the sound waves. Hence, since H is proportional to the current in the coils, this current will be directly proportional to the intensity of the sound. Since relative loudspeaker efficiencies were the chief concern in this work, these torques are here expressed in terms of this current in milliamperes.

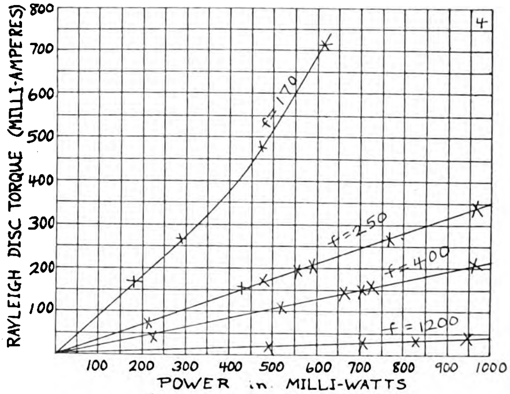

Fig. 7—Sound Intensity Compared to Power Input to the Loudspeaker at Various Frequencies.

To determine zero position for the disc, light sent through the window and back from the mirror mentioned in a preceding paragraph was focused at the source and received on a direct vision screen.

Since the operation of the Rayleigh disc requires considerable care, all the sound intensity measurements were made at one time and the data at each frequency plotted against electrical input to the loudspeaker. Fig. 7 shows the calibration curves for the various frequencies used. With the aid of these curves the sound intensity could be determined for each setting of an observer by simply reading the electrical power input.

Wattmeter

For the purpose of measuring the power input to the loudspeaker a wattmeter of the electrodynamometer type was constructed. The dynamometer with the front of the case removed

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may be seen in Fig. 8, and a wiring diagram showing internal and external connections of the instrument is shown in Fig. 9. It will be seen that the instrument functions as a simple indicating wattmeter.

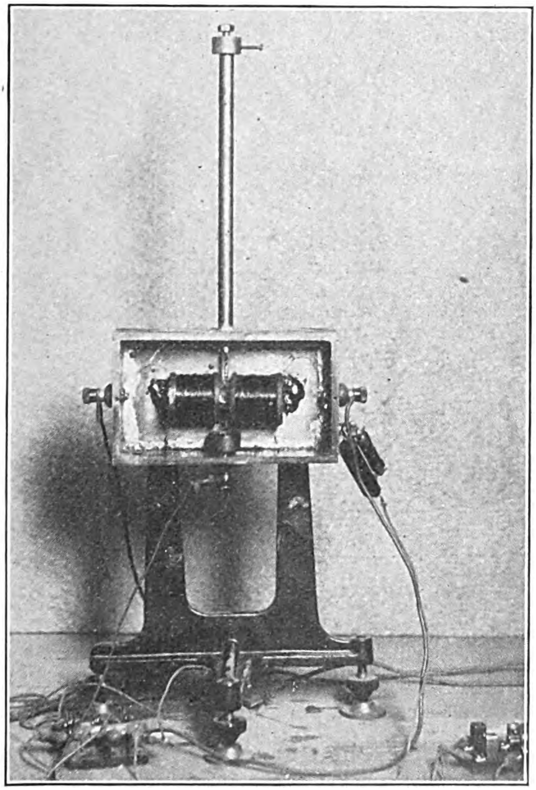

Fig. 8—The Electro-Dynamometer with the Front of the Brass Case Removed.

The moving coil of the wattmeter consists of 2,100 turns of No. 40 enameled cupron wire wrapped on a fiber form of about 2.0 cm diameter. A resistance of 50,000 ohms is in series with this coil. The coil resistance is 3740 ohms giving a total of 53,740 ohms for the potential circuit of the dynamometer. The stationary coil is divided into two parts with the movable coil between them as shown in the photograph. The stationary coil is made of 1,200 turns of No. 28 enameled copper wire. Its resistance is 37 ohms.

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The movable coil has a standard galvanometer suspension of phosphor bronze ribbon above and a coil of phosphor bronze ribbon beneath itself. This coil makes contact through a small copper container which carries oil. Two wire prongs extend downward from the moving coil into this oil and serve to damp the oscillations of the coil. A galvanometer mirror is also mounted directly on the moving coil.

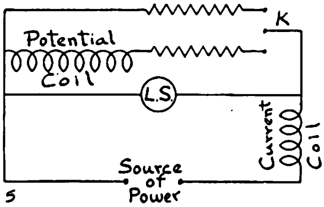

Fig. 9—Schematic Diagram of the Wattmeter Showing Connections to the Source of Power and to the Loudspeaker.

To avoid any errors that might arise from lack of shielding, frequency variation, energy used by the dynamometer itself, etc., the instrument was calibrated in place at every frequency and throughout the range of power used. Fig. 10 shows the

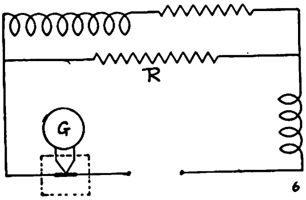

Fig. 10—Schematic Diagram for the Calibration of the Wattmeter.

arrangement used for calibrating. Power was supplied at the various frequencies to a pure resistance. The current was measured by means of a Western Electric thermocouple calibrated with direct current and a milliameter.

The deflections of the wattmeter are read by means of a telescope and scale placed at a distance of 200 cm. Fig. 11 shows a calibration curve for 25 cycles, 250 cycles, and 1,200 cycles. Table I shows the necessary data for the use of the dynamometer at all the frequencies employed.

TABLE I

Sensitivity of Wattmeter

Frequency	Milliwatts per cm.
25	67.0
170	68.5
250	70.4
400	73.0
1,200	77.3

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While the alternation phonometer was operating and an observer was adjusting the intensity of one of the tones the switch K shown in Fig. 9 was turned so as to cut the potential coil out of the circuit and put a conductor of approximately equal resistance in its place. As soon as the observer decided on his adjustment, the rotating condensers were stopped with either C_1 or C_2 in its maximum position. Energy to the other condenser was cut off by removing the plate potential from the oscillator-amplifier feeding it. The potential coil of the dynamometer was then placed in the circuit by means of switch K

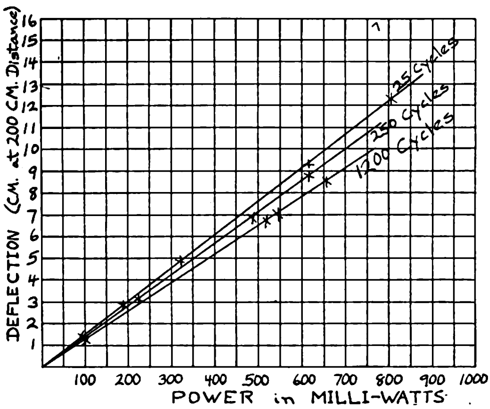

Fig. 11—Calibration Curves for the Wattmeter at Various Frequencies.

and the deflection of the dynamometer determined. The condensers were then rotated through 180 deg. and the power input to the loudspeaker from the other source determined in a similar manner. The frequency of the variable source was then altered and the process repeated. It was of course necessary to keep the intensity of the tone used as a basis for comparison as nearly constant as possible.

Experimental Results

Fig. 12 shows a single set of observations of sound intensity versus frequency for equal apparent intensity of tones using a frequency of 400 at an electrical input of 755 milliwatts as a base. This data is represented by the crosses. The circles show the average of 3 sets of data taking a frequency of 1,200 at an electrical input of 1055 milliwatts as a base. The squares are the average of 2 sets of data using a frequency of 250 at 332 milli-

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watts as a base. The solid line shows the average of 4 sets of data using each tone once as a base. The conformity of the curves for all the observations shown in this figure to the same general shape is an indication of the accuracy of judging equal apparent intensities with the aid of the alternation phonometer described even at rather high loudspeaker intensities.

In Fig. 13 we have a record of the same observations as in the solid curve of Fig. 12 except that here we have plotted power input against frequency. This figure is of course not of general interest, since it applies only to the loudspeaker used in these tests.

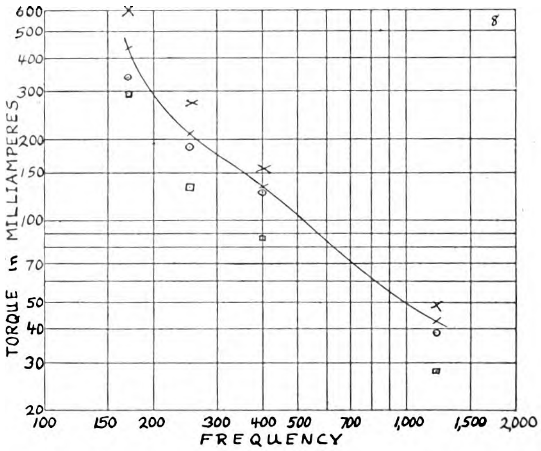

Fig. 12—Observations of Sound Intensity Versus Frequency for Equal Apparent Intensity. (See text for explanation of marks.)

These curves (Fig. 12) show, as was to be expected from the work of previous investigators, that the ear sensitivity falls off for the lower tones. On the other hand, we see by plotting the ratio of Rayleigh disc torques to electrical input against frequency for equal intensity (data for the observations shown in the solid curve of Fig. 12) that the loudspeaker used is more efficient in our lower range than in the upper. (See Fig. 14.) A similar curve plotted for equal actual intensities would show this characteristic still more.

Except for a frequency of 170, Fig. 7 indicates that the efficiency of this loudspeaker does not vary with intensity at a given frequency in the region covered by this graph.

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Sources of Error and Limitations of the Device

The alternation phonometer did not prove as satisfactory as a manual switching of sources when both tones were at approximately the same frequency, due largely to beat notes.

The plates of condensers C_1 and C_2 were semicircular. It will be readily seen that three questionable things occur with commutation by means of these condensers. First, at no time is

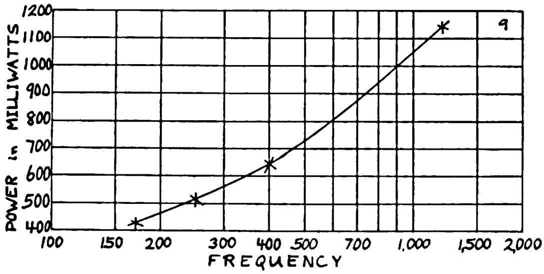

Fig. 13—Power Input Versus Frequency for Equal Apparent Intensity

the capacity of either condenser zero, and consequently at all times both tones are present to some extent. Second, the intensity of either tone approaches a maximum and at once starts to recede. It does not stay at constant intensity even approximately for any large percentage of the time. Third, twice during every cycle of the alternator both tones are present simultaneously at the same intensity. Whether or not any or all of these items are seriously objectionable is difficult to say. The second item could probably be rectified by using condenser plates of different shape.

It is probable that the absolute values of sound energy as could be determined from the Rayleigh disc torques might be in error, in the first place, due to another torque from radiation pressure. The possibility of error from this cause can be obviated by repeating the measurements with the disc rotated through 90 deg. and the two sets of data averaged.[6] It was not thought necessary to take this precaution for the purposes of this experiment.

There is also the possibility of error due to reflections of sound waves from the walls of the compartment. Essentially this phenomenon may be thought of as sound waves approaching

[6] S. L. Quimby, "On the Experimental Determination of the Viscosity of Vibrating Solids," *Phys. Rev.*, 25, 558, 1925.

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the disc from various directions. The result of such reflections may be either to increase or decrease the torque on the disc. Judging from the data of Watson,[7] it seems unlikely that much more than 25 per cent of the incident energy could be reflected. From the general nature of the Rayleigh disc calibration curves for various frequencies shown in Fig. 7 it would seem that no very prominent standing wave patterns existed in the sound compartment. Had they existed, the disc would of necessity have found itself in different parts of the pattern for different

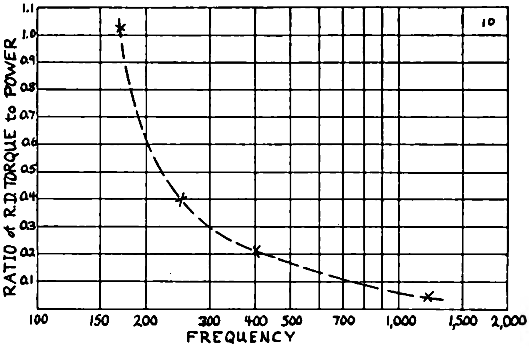

Fig. 14—Apparent Relative Efficiency of Loudspeaker.

frequencies. These curves do not indicate such an effect. It would seem that the value of the torques on the Rayleigh disc are at least dependable for relative values of intensity. We hope, however, to improve this feature for further measurements.

The principal source of error is probably traceable to the use of too small a power tube to feed the loudspeaker. This matter will also be taken care of in future measurements.

Utility of the Alternation Phonometer

In some respects this paper is in the nature of a preliminary report. However, its publication at this time seems warranted by the expectation that it describes a method applicable to loudspeaker measurements in laboratories with limited facilities for such measurements.

In many cases of practical work with loudspeakers the matter of relative efficiency at various frequencies is of more importance than actual efficiency of the speaker. It is therefore hoped that the work here described may suggest the possibility of making

[7] F. R. Watson, *Phys. Rev.*, 7, 125, 1916.

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analyses of loudspeakers over a considerable frequency range in commercial work. For such work the Rayleigh disc could be omitted and only the wattmeter and the alternation phonometer used. For rough work the sound compartment could be omitted, but for more nearly reproducible results it will be found quite necessary.

Acknowledgment

The authors are indebted to Mr. L. C. F. Horle, of the Federal Radio Corporation, for advice and suggestions during the progress of this work.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

FACSIMILE PICTURE TRANSMISSION*

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Summary—*A facsimile picture transmitting system is described. The chief object of the design of this system was to produce a simple, rugged apparatus for practical usage, which would not require the attention of a skilled operator. The system does not require a special preparation of the original, and the receiver records the copy directly on the photographic paper.*

The usually delicate problem of photo-cell current amplification has been simplified to such an extent that only three stages of resistance coupled amplification suffice between the photo-cell and modulator of the broadcasting station. This was made possible through the design of a very efficient optical system, which supplies to the photo-cell quite enough light reflected from the picture even though only a small incandescent lamp for illumination is used.

The synchronizing and framing have also been simplified to such a degree that they do not require any special channels or special amplifiers.

Automatic starting devices obviate the use of any complicated scheme of signal dispatch for starting the apparatus. In spite of the simplicity of operation, it is capable of transmitting a 5 in. by 8 in. picture either in black and white or in half-tone in 48 seconds, or a message at the rate of 630 words per minute over short distances.

The resulting picture prints are of a quality quite satisfactory for newspaper reproduction and clear facsimile of messages may be made from typewritten originals.

OPTICAL SYSTEM

ALL the existing methods of electrical picture transmission can be divided into two classes: one which requires special preparation of the original before it can be transmitted, and the other which can transmit the original directly.

The first one includes the electrical contact method, now almost abandoned, which requires the preparation of the original in such a form that the dark and light of the picture give variable electrical resistance when explored by a traveling contact.

In another form of transmitter of the same class, as in the Belin system, which is still in commercial use in France, the picture is embossed with a special ink so that it may be reproduced by a microphone in the same way as an electrical pick-up reproduces phonograph records.

The method which requires the preparation of a transparent picture either in negative or in positive form belongs also to

* Dewey decimal classification: R582. Original manuscript received by the Institute, November 24, 1928. Presented before New York meeting of the Institute, January 2, 1929.

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the same class.[1] In this case the picture is explored by a sharply defined pencil of light which passes through it and activates a photo-cell placed behind it. The variation in optical density produces a corresponding variation in absorption of the light, and therefore the photo-cell delivers an electric current varying according to the picture.

The present requirement of high-speed transmission, however, rules out all these methods due to the time necessary for preparation of specially treated originals. In this case only one solution remains, and this is the scanning of the original by a pencil of light and the utilization of the light reflected from its surface. The amount of reflected light is directly dependent on the density of the picture or lettering. The specular reflection

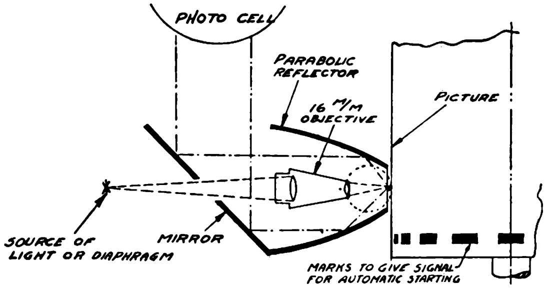

Fig. 1—Optical System of the Picture Machine.

from the surface of the paper is negligible and equal at all points on the paper, and therefore does not interfere with the reflecting scanning method. However, difficulties arise in the optical part of the problem due to the small amount of light reflected. This necessitates the utmost care in the design of a very efficient optical system.

Here again are two possible solutions of the problem. One is the illumination of the original by strong diffused or concentrated light and cutting off from the illuminated area a small, sharply defined spot. The light from this spot is directed into a photo-cell. In this case the optical efficiency is determined first by the ratio of the size of the scanning spot to that of the total illuminated area, and by the light-gathering power of the optical lens. In general, in spite of all the precautions, the over-all optical efficiency of this method is quite low.

[1] This method is still in general use and, a few years ago, was the only optical method employed.

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The second method, which is used in this transmitter, is inverse to the first one. In this case the size of the scanning spot is adjusted to the required dimension and the reflected light is collected. Fig. 1 gives an idea of the arrangement. The source

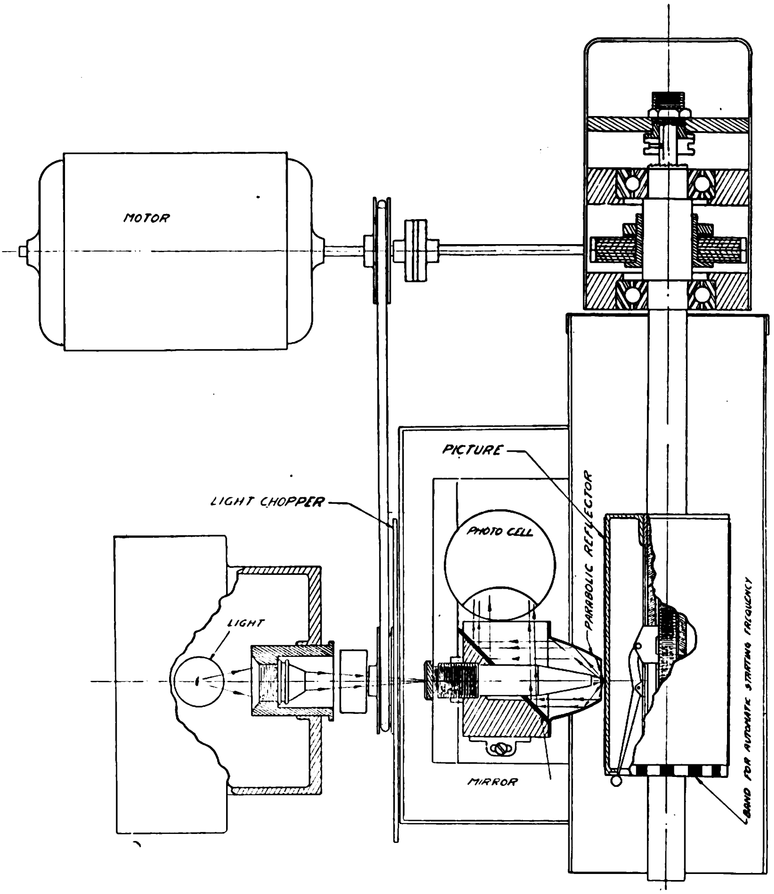

Fig. 2—Facsimile Picture Transmitter.

of light is focused first on a diaphragm to make the size of the spot independent of the size of the source. The image of the diaphragm, with necessary reduction, is focused on the surface of the picture. The reflected light is gathered by means of the parabolic reflector, the focus of which coincides with the illuminating point. Part of the reflector is cut away in order to pass the light, and the remaining part is brought into close proximity

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to the surface of the picture. In this case almost all the reflected light is collected and projected as a more or less parallel beam by the reflector. A plane mirror with a small hole for passage of the illuminating spot intercepts the reflected light at 45 degrees and diverts it to the photo-cell. The optical path of the whole system is quite short and the construction is flexible for adaption to almost all kinds of scanning systems. The over-all optical

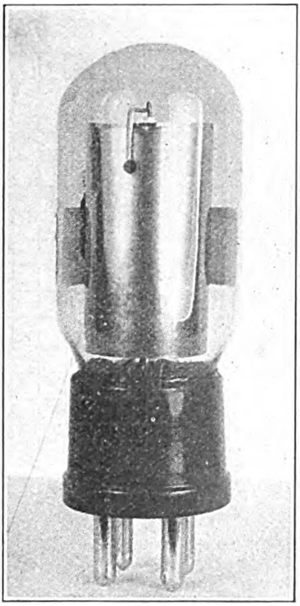

Fig. 3—Glow Tube.

efficiency is many times greater than the best possible solution by the first method.

Scanning Arrangement

In the present article only the intermittent type transmitter, i.e., the type in which it is necessary to stop the machine after every picture for reloading, is described. Although this type is not very suitable for commercial purposes, it has many advantages for experimental work and also for all kinds of communication where traffic requirements are not very high.

In this type, the picture to be sent is in the shape of a 5 in. by 8 in. rectangle and is wrapped around a cylinder which is placed on the shaft as shown in Fig. 2. The shaft is hollow and

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has a screw inside which can be locked either to the shaft or to the support. The cylinder has a locking nut, which, by means of a lever, can be raised or lowered, locking the cylinder with the shaft and screw. While the screw is connected to the shaft, the cylinder rotates with it, but remains stationary in longitudinal direction. When the picture is ready for transmission, the screw is released from the shaft and locked to the support by means of an electromagnetic clutch. This makes the screw stationary in respect to the rotating nut and the cylinder begins to advance,

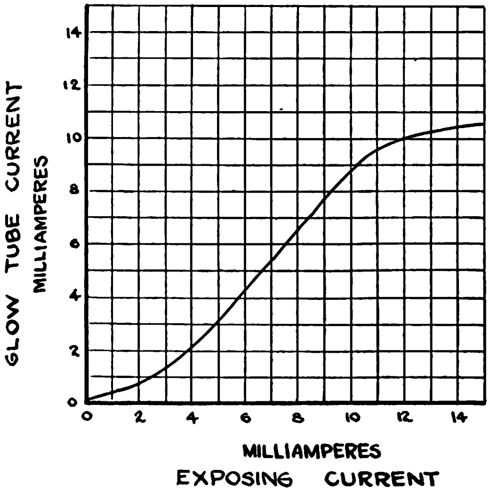

Fig. 4—Fidelity Curve.

exposing gradually the whole surface to the scanning spot. The speed of scanning is 56 in. per second, while that of the cylinder feed is 1/64 in. per revolution, so that 52.5 sq. in. of the picture are covered per minute. Allowing, as usual, 12 words per sq. in., the rate of transmission amounts to 630 words per minute. The 1/64 in. picture feed was found very satisfactory for type-written messages and for most other pictures.

Care, of course, is taken to move the cylinder with uniform speed in order to avoid distortion of the picture.

Mechanically, the receiver is identical with the transmitter. A standard bromide photographic paper 5 in. by 8 in. size is wrapped around the cylinder and the recording is done by means of a glow discharge tube. Fig. 3 shows the type of glow tube used for this purpose. It was developed by Mr. Knowles in the

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Westinghouse Laboratory and uses helium glow for recording. The glow is restricted to the required size by a mask built into the tube. A discharge of approximately 15 milliamperes at 400 volts is sufficient at the present speed to produce very satisfactory blackening on the bromide paper.

Fig. 4 shows the fidelity curve for the whole transmitting process. Along the axis of abscissas is plotted the current through the glow tube. The bromide paper is exposed to this glow at working speed and is developed in the usual manner. The "density chart" prepared in this manner is put into the trans-

Fig. 5—Photo Cell.

mitter, and the output of the amplifier is plotted along the axis of ordinates. The curve, therefore, represents the true relation between the densities of the original and of the reproduced copy.

This relation shows good proportionality from 2 to 12 milliamperes, which ratio usually is maintained for transmission of half tone pictures. For white and black manuscripts it is preferable to sacrifice the proportionality and work further along the characteristic curve in order to produce sharper contrast.

For a highly efficient commercial transmitter, the intermittent type is not suitable, due to loss of time required for reloading. Another machine of continuous type has been developed for this purpose. Although the principle and even the optical system remains the same, the relative motion of the paper with respect to the scanning light is changed, and this

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in turn changes the whole appearance of the apparatus. The reflected light is conveyed by plane mirrors along the axis into a photo-cell, which remains stationary.

Photo Cell

Fig. 5 shows a photo-cell used in the picture transmitter. It is of magnesium-caesium type, filled with argon. It consists of two electrodes, one on the inner surface of the glass bulb and another in the shape of a ring in the center of the bulb. A window is provided in the coating for admittance of the light. The coating is photo-sensitive; i.e., it emits electrons at a rate

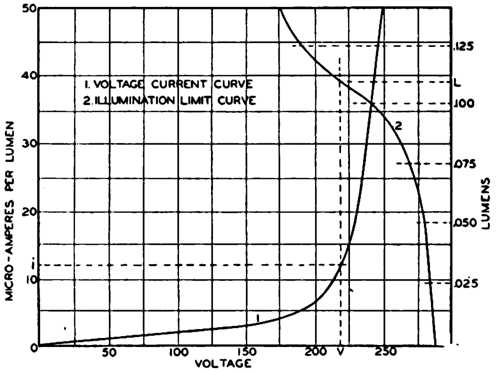

Fig. 6—Characteristics of Gas-Filled Photo-Electric Cell.

proportional to the quantity of light absorbed by the coating. These electrons flow to the anode under the accelerating potential of an outside battery. During this passage they collide with the molecules of the argon, and since their velocity-voltage is higher than the ionizing potential of the argon, ionization occurs. Thus the output of the cell is increased many times without destroying the proportionality between the quantity of light absorbed by the photo-cell and the output of the cell. Fig. 6 shows the relation between the voltage applied to the cell and its output calculated in microamperes per lumen. The second curve on the same figure gives an idea of the safe operation of the cell for various voltages and degrees of illumination.

Amplifier

Since the photo-cell, under operating conditions, supplies a current of the order of 1/20 of a microampere for the white

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portion of the picture, a strong amplification is necessary before the output of the cell can be used for radio transmission.

The requirements for the amplifier are quite severe. It should not distort the signals and should not have a tendency to oscillate, which results in distortion of the picture.

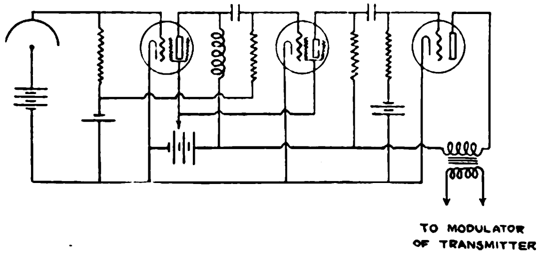

Fig. 7—Photo-Cell Amplifier.

In actual cases we used screen-grid tubes and the circuit as shown in Fig. 7. Voltage output of the third tube is about 40 volts, which is quite sufficient to operate the modulator of the broadcasting station through a line of considerable length. Fig. 8 shows an oscillograph actually received from the picture signals.

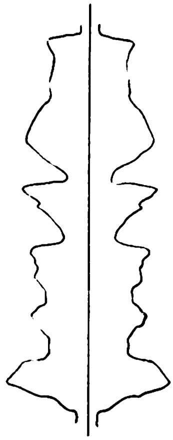

Fig. 8—Envelope of Signal Current.

Receiver and Amplifier

For the reception of the picture signals, the radio set may be a standard receiver. The amplification, in the case of weak signals, is preferably at radio frequency in so far as possible, in order to reduce distortion. Transformer-coupled audio-frequency amplification, however, gives very good results if the gain is

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fairly uniform between 2000 and 4000 cycles. To date, a standard R.C.A. short-wave receiver has been used for all work. This employs a stage of radio-frequency amplification with a screen-grid tube, detector, and two audio stages.

Control of Glow Tube

For the control of the glow tube which exposes the photographic paper, a vacuum tube is used. The glow tube is connected in series with the plate voltage for the vacuum tube. Fluctuations in voltage upon the grid due to the picture signal produce corresponding variations in the glow tube current. If the signal as it comes from the audio amplifier were applied

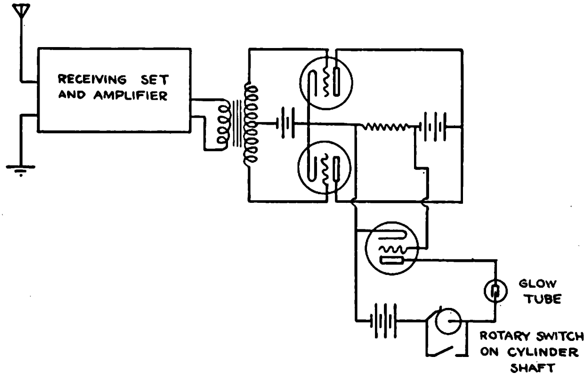

Fig. 9—Glow Tube Control Circuit.

directly to the grid of the control tube, a negative picture would result. That this is true can be verified by following the steps in the transmission of the picture. When the light is reflected from a white area in the original picture, a relatively large amount reaches the photo-cell. The corresponding photo-cell current is amplified, producing a loud signal. At the receiving end this signal would increase the average plate current of a tube working on the lower bend of the characteristic curve. Such an increase would augment the light from the glow tube, darkening the photographic paper instead of making it lighter.

Unless the picture is to be recorded upon a film, and subsequent prints are to be made, it is necessary to reverse the process. This reversal might take place at the transmitter, but is undesirable for pictures which are largely white, as printing, for example. Bursts of static would be recorded as black spots

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on the white background. On the other hand, if the reversal occurs at the receiver, these disturbances tend only to make the white whiter. Hence, reversal at the receiver is used.

The circuit employed for the control of the glow tube is shown in Fig. 9. Voltage from the receiving set is applied to the push-pull detector, using UX-112 tubes. These are so biased as to give practically zero plate current in the absence of signal. Any voltage supplied causes, on either the positive or negative half of the cycle, a voltage drop across the plate resistor. This

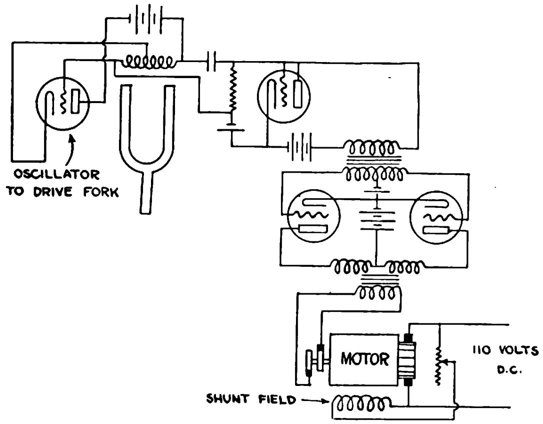

Fig. 10—Synchronizing System.

voltage drop is impressed on the grid of the control tube, decreasing the glow tube current in the case of a strong picture signal.

Synchronizing

The problem of synchronizing transmitter and receiver is of great importance, particularly for high speed transmission. The plan of broadcasting a standard frequency by a series of stations scattered throughout the world is one of considerable merit, but has not yet been adopted. The use of voltage from interconnected power lines has been proposed, but is impractical in the general case, since the phase relations between ends of the system are too variable to permit high speed transmission. It is necessary, then, to do one of three things: (1) provide independent, accurate sources of frequency for transmitter and receiver, or (2) send a synchronizing signal continuously to the receiver, or (3) correct periodically a less accurate source of

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frequency at the receiver by an impulse from the transmitter. The third method is the one used.

The source of frequency at both transmitter and receiver is a 70-cycle tuning fork in a constant temperature box. These forks are so adjusted that there is but one beat between them in 20 seconds or more; this condition is relatively easy to maintain. The fork at the receiving machine is then corrected every revolution of the picture cylinder (every seventh of a second) by an impulse of about one-half cycle duration. This impulse is transmitted over the same channel as the picture, but on the margin of the paper to avoid interference with picture signals.

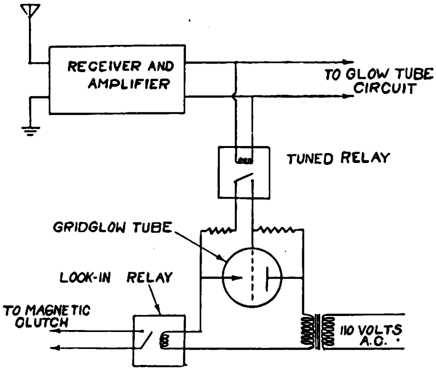

Fig. 11—Automatic Starting Arrangement.

Having obtained the standard frequency, the next step is to use it most advantageously in the control of the motors. To amplify a small amount of energy to such a degree that it could supply the full load of the machine would be wasteful. It is common practice at the present time to use two motors on the same shaft—one to furnish most of the torque, and one to keep the speed constant. In the present arrangement, the two machines are combined into one, similar to a rotary converter. Voltage from the tuning fork is amplified, using two UX-250 tubes in the final stage. The power from these tubes is applied to the motor slip rings, while most of the energy comes from the direct-current source. Fig. 10 shows the schematic diagram of the synchronizing circuit.

Framing of Picture

It is not enough that the cylinders on both transmitter and receiver rotate at exactly the same speed; the picture must be

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framed as well. In other words, the first edge of the picture being sent should be under the spot of light at the transmitter at the same instant that the first edge of the photographic paper is being exposed by the glow tube at the receiver.

The framing is accomplished by the following method. The picture to be transmitted is held on the cylinder by a longitudinal

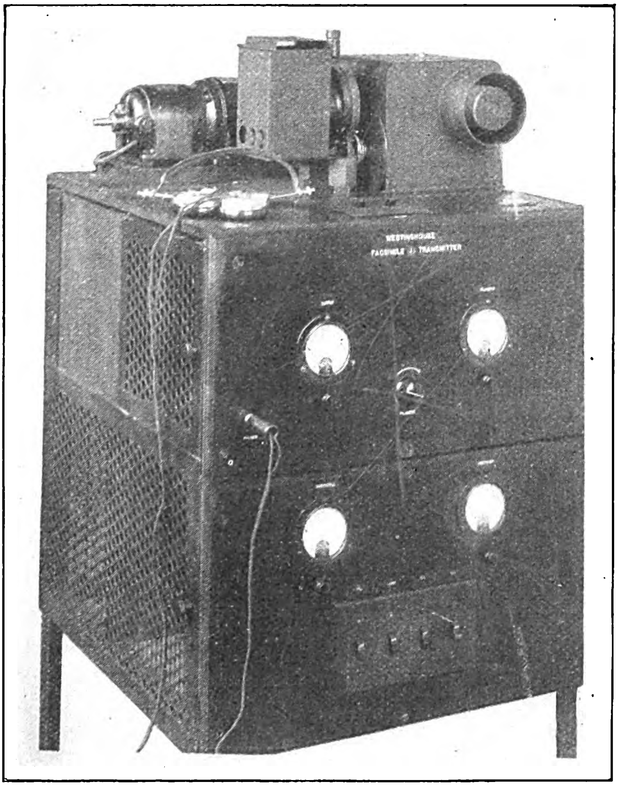

Fig. 12

black band; at the end of the cylinder first transmitted is a narrow white ring. As the light spot explores this ring, a continuous signal is transmitted except for the interval when the black band is absorbing most of the light. The glow tube at the receiver flashes once for each time the black band occurs, or seven times per second. At the end of the shaft upon which the receiving cylinder rotates is an interrupter which breaks the glow tube current for a time equal to that required for the trans-

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mission of the band. If the interruption takes place at the same time the flash normally occurs, the light from the glow tube appears steady, and framing is known to be correct. If the flashes are seen, however, it is necessary to correct the relative position of the glow tube with respect to the position of the cylinder at a given instant. This is done by a process equivalent to rotating the frame of the motor.

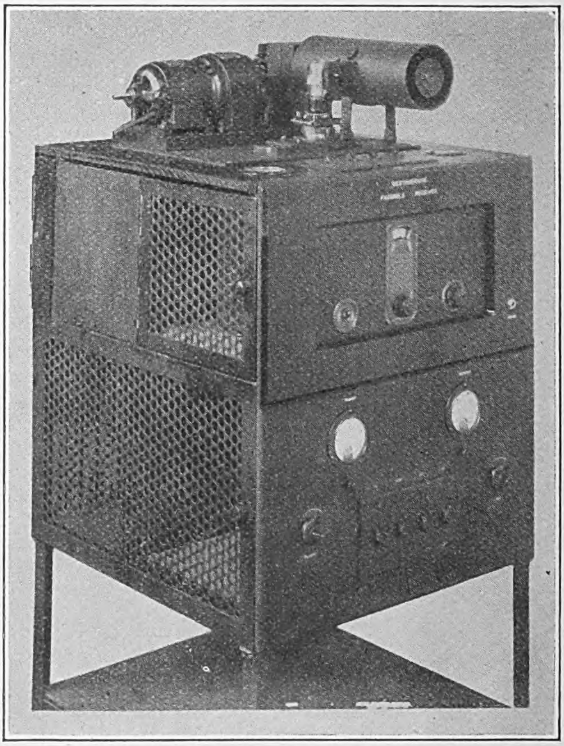

Fig. 13

The framing process described above is carried out for each picture transmitted by the intermittent machine, since the framing is lost when the motors are stopped. In the continuous machine the motors run constantly, hence framing is required only at the beginning of transmission.

STARTING OF THE RECEIVER

When synchronizing and framing is accomplished the picture starts to pass under the transmitter's scanning spot. The

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starting of the receiving cylinder is accomplished automatically. The principle of operation of this starter is as follows: On the front end of the transmitting cylinder a band of black and white spots is engraved. This can be seen on Fig. 1. When the picture is started, this band comes first under the scanning spot. As a result, the corresponding frequency is produced by the transmitter and reproduced by the receiving amplifier. This frequency operates a small tuned relay, Fig. 11, which in turn starts a grid glow tube.[2] The current passing through the grid glow tube operates a lock-in relay which completes the circuit

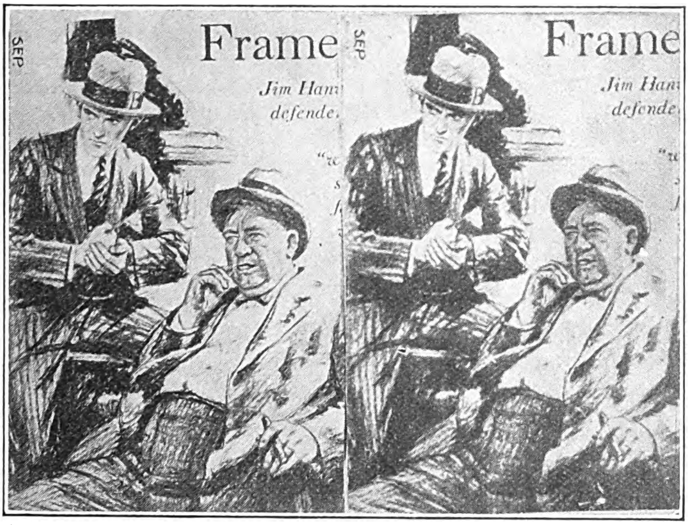

Fig. 14

to the magnetic clutch; this starts the receiving cylinder. The starter does not require any additional equipment at the transmitter, nor at the broadcasting station.

Assembled Machines and Results Obtained

Figs. 12 and 13 show the finished appearance of the intermittent type of transmitter and receiver, respectively. Both machines are self-contained, including all the amplifiers, rectifiers, and tuning forks. In size, the cabinets are two ft. square by four ft. high. The transmitter could be installed at any convenient place connected with the broadcasting station by

[2] See description of grid glow tube by Knowles, *Elec. Jour.*, April, 1928.

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means of a telephone line. The receiver should be placed either in a dark room or adjacent to a small developing booth. With the exception of the darkening of the end of the receiving machine for handling the bromide paper, no other precautions are required in the illumination of rooms where both machines are located. In Fig. 14 are shown side by side an original picture and the facsimile transmitted over a short telephone line and a few miles of radio channel.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

NOTES ON GRID-CIRCUIT DETECTION*

By
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Summary—*A dynamic method of finding $\partial^2 i_g/\partial e_g^2$ or $\partial k_g/\partial e_g$, the main term in grid-circuit rectification, is given. This method is based upon formulas given by Chaffee and Browning*[1] *and consists in calculating $\partial k_g/\partial e_g$ from the change of d.c. when a known a.c. input voltage is applied to the grid. The values of $\partial k_g/\partial e_g$ found by the dynamic method are compared with the values found by the usual method. The effect of frequency, internal grid resistance, and external resistance on the detection voltage introduced in the plate circuit are also considered. An experimentally determined curve is given showing the detector frequency distortion of a commercial set for a 2-megohm grid leak and for a ½-megohm grid leak.*

THE subject of detection has been covered from the theoretical side in two excellent articles. The article by Chaffee and Browning[1] gave a comprehensive theoretical treatment of detection together with a bridge method of experimentally determining the detection coefficient. The article by Llewellyn[2] covered the general theory of vacuum-tube operation. The theory in the above article was complete and the equations were applied in some detail to grid-leak detectors. An article by Smith[3] gave a good physical picture of grid rectification as far as the grid circuit itself was concerned.

The main difficulty in using the theoretical analysis as given by the first two mentioned articles is in determining the second derivatives of the currents with respect to the voltages over the range of input voltages applied to the detector. To determine these derivatives it is necessary to plot the conductance curve from the static characteristic curve, then plot another curve from the conductance curve. There is also the uncertainty of whether the dynamic value will correspond with the static value found graphically. If the values of the second derivatives

* Dewey decimal classification: R134. Original manuscript received by the Institute, November 14, 1928.

[1] E. L. Chaffee and G. H. Browning, "A Theoretical and Experimental Investigation of Detection for Small Signals," Proc. I. R. E., **15**, 113; February, 1927.

[2] F. B. Llewellyn, "Operation of Thermionic Vacuum Tube Circuits," *Bell System Tech. Jour.* **5**, 433; July, 1926.

[3] Lloyd P. Smith, "Theory of Detection in a High Vacuum Thermionic Tube," Proc. I. R. E., **14**, 649; October, 1926.

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could be found by some simple means for the desired value of input voltage the usefulness of the theory would be increased.

The theory itself could be checked and the range over which the theory holds could also be found. Different types of tubes could be compared easily as the main terms in both plate and grid rectification are the second derivatives of the currents with respect to the voltages.

It is the purpose of this article to show that the value of the second derivative of the grid current with respect to the grid voltage, the main term in grid rectification, may be found by the change in direct current in the grid circuit with any de-

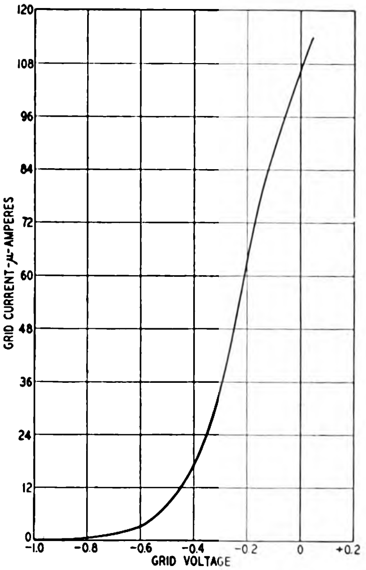

Fig. 1—Grid-Current Grid-Voltage Characteristics; Typical C-327 Tube.
Eb—45 v.
Ef—2.5 v.

sired value of input voltage. The results are also a check on the theory. In another article the author[4] shows how the value of the second derivative of the plate current with respect to the grid voltage, the main term in bias detection, may be found when the input voltage is fairly large.

A sensitive direct-current meter with a full scale reading of about five microamperes is required. Good accuracy may

[4] J. R. Nelson, "Detection with the Four Element Tube," Proc. I. R. E., **16**, 822; June, 1928.

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be obtained by balancing out the direct current if it is more than three or four microamperes. There is no difficulty in using a meter of this type in the grid circuit, but it is difficult to keep the plate current constant enough to use it in the plate circuit of a three-element tube. Chaffee and Browning[5] give the following formulas for the increment of direct current.

$$\overline{\Delta^2 I_g} = \frac{\overline{\Delta^2 E_g}}{r_g} + \frac{1}{4}(\Delta E_g)^2 \quad \frac{\partial K_g}{\partial e_g} \tag{1}$$

Where

$\overline{\Delta^2 I_g}$ = The increment of direct current
$\overline{\Delta^2 E_g}$ = The increment of steady voltage
$\underline{\Delta E_g}$ = The maximum value of input voltage
r_g = Grid resistance
K_g = Grid conductance
$\overline{R}$ = Resistance to a steady current

$$\overline{\Delta^2 E_g} = -\overline{R}\Delta^2 I_g \tag{2}$$

$$\overline{\Delta^2 I_g} \times (r_g \overline{R}) = \frac{r_g}{4} \frac{\partial K_g}{\partial e_g} \underline{(\Delta E_g)^2} \tag{3}$$

$$\overline{\Delta^2 I_g} = \frac{r_g}{r_g + \overline{R}} \frac{1}{4} \underline{(\Delta E_g)^2} \frac{\partial K_g}{\partial e_g} \tag{4}$$

$$\frac{\partial K_g}{\partial e_g} = \frac{4\overline{\Delta^2 I_g}}{\underline{(\Delta E_g)}^2} \frac{r_g + \overline{R}}{r_g} . \tag{5}$$

If the d.c. resistance is small compared to r_g Eq. (5) reduces to

$$\frac{\partial K_g}{\partial e_g} = 4 \frac{\overline{\Delta^2 I_g}}{\underline{(\Delta^2 E_g)}^2} . \tag{6}$$

Fig. 1 shows the grid-current grid-voltage characteristic of a typical C-327 tube with $Eb = 45$ volts and $Ef = 2.5$ volts. The curve is steeper than the curves obtained with other types of tubes such as the CX-301 A, CX-340, etc. The tangents are difficult to obtain accurately, but it was felt that if the theory could be checked using this type of tube the method of determining $\partial K_g / \partial e_g$ by the change in direct current could be applied to other types of tubes without the necessity of checking the values graphically.

[5] Loc. cit., page 129.

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Fig. 2 shows the grid conductance, $\partial i_g/\partial e_g$ of the typical C-327 tube, plotted against the grid voltage over the range which is of most interest. This curve was obtained from Fig. 1 by finding the tangent $\Delta i_g/\Delta e_g$ at the different values of grid voltages, then plotting the values of tangents against the grid voltages. Fig. 3A shows $\partial K_g/\partial e_g$ or $\partial^2 i_g/\partial e_g^2$ plotted against the grid voltage. This curve was determined from Fig. 2 by finding the values of $\Delta K_g/\Delta e_g$ at the different grid voltages.

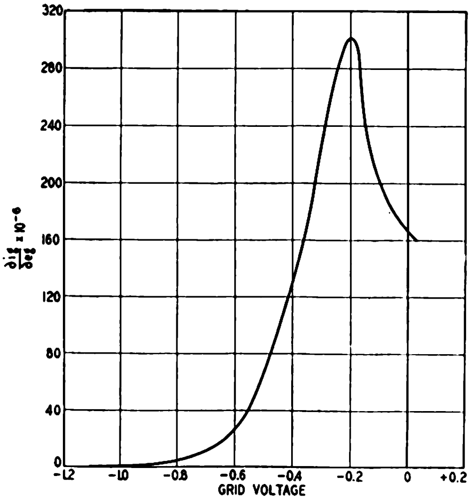

Fig. 2—Grid Conductance, $\partial i_g/\partial e_g$, of Typical C-327 Tube.
Eb—45 v.
Ef—2.5 v.

This is the usual method of determining the value of $\partial K_g/\partial e_g$ or $\partial^2 i_g/\partial e_g^2$. If the values of $\partial K_g/\partial e_g$ found from the change of direct current as explained below check the values of $\partial K_g/\partial e_g$ found from the static characteristic as explained above, the assumptions made in developing the theory are justified.

Fig. 4 shows the circuit used to determine the values of $\partial K_g/\partial e_g$ at different operating voltages. Alternating current from the source flows through the meter A_2 and the resistance R. The value of R was less than 10 ohms so that the input voltage is equal to the product of the reading of A_2 and R. No load is shown in the plate of the C-327, but if desired the normal load may be placed in the plate circuit. The a. c. source may be either radio frequency or audio frequency, as the theory is independent of the frequency of $\underline{\Delta E_g}$. If a specific case is be-

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ing investigated by placing a load in the plate circuit the input voltage should be radio frequency so as to take into account the input impedance of the detector at radio frequency.

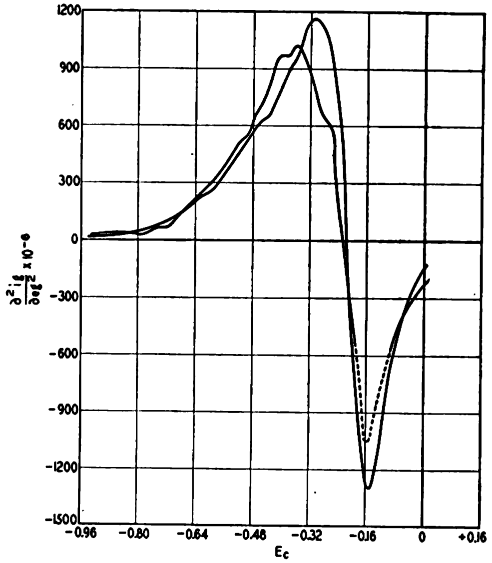

Fig. 3—Variation of Grid Conductance, $\partial K_g/\partial e_g$, of Typical C-327 Tube.
Eb—15 v.
Ef—2.5 v.

The meter A_1 had a resistance of about 5000 ohms so that it was necessary to use Eq. (5). The value of r_g was determined from Fig. 2. The meter A_1 was shorted for a.c. with a large con-

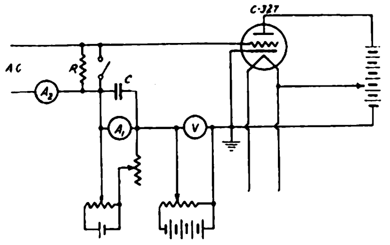

Fig. 4

denser so that it was not necessary to correct for ΔE_g. After the grid current was above five microamperes it was balanced out. The resistance of the balancing out circuit was high, so

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that it was not necessary to correct the increment of d.c. as read on A_1.

The d.c. was balanced out of A_1 and R was shorted out. The peak volume of a.c. across R was kept constant at 0.035 volt. The short circuiting switch around R was then opened, and the increment of d.c. was read on A_1. This was repeated for different values of E_c.

The values of $\partial^2 i_g/\partial e_g^2$ were then calculated from Eq. (5) for the different values of E_c. The values thus found are shown plotted in Fig. 3B. The agreement is good considering the difficulty of finding Fig. 3B. The peaks are separated less than the total grid swing, $2\times 0.035 = 0.07$ volt. The dotted portion of B is estimated, as no accurate voltage between -0.1 and -0.2 of a volt could be determined with the one-volt thermo-voltmeter used.

The curves of Fig. 3 do not show where the point of maximum detector sensitivity will occur. They show only the value of voltage to which the grid should be biased to obtain the maximum current when the external grid circuit is shorted. In determining the point of maximum voltage on the grid the values of the internal and external grid impedances as well as the value of $\partial K_g/\partial e_g$ have to be taken into account.

Chaffee and Browning[6] give the following equation for the voltage introduced in the plate circuit.

$$(\text{Det. E})_l = \frac{r_p}{4}\left[-\frac{R_1}{\sqrt{R_1^2C_1^2w^2+1+R_1K_g(2+R_1K_g)}}\,g_m\frac{\partial K_g}{\partial e_g} + \frac{\partial g_m}{\partial e_g}\right] B\sqrt{2}m(\underline{\Delta E_0})^2 \qquad (7)$$

where

$(\text{Det. E})_l$ is the audio-frequency voltage introduced in the plate circuit.

R_1 is the value of the external grid resistor.

C_1 is the value of the external grid condenser.

Eq. (7) neglects the effect of the input conductance not due to the electron flow and the input capacity of the tube at

[6] Loc. cit., page 142. Note: If the part due to bias rectification is neglected the same equation may be obtained from Eq. (43), page 448 of the article by F. B. Llewellyn, "Operation of Thermionic Vacuum Tube Circuits," *Bell System Tech. Jour.* 5, No. 3; July, 1926.

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audio frequencies. If it is desired to take the above factors into account the article of Ballantine[7] may be consulted.

The voltage introduced in the plate circuit given by (7) may be used as μe_g in the usual amplifier equations, and the value of the voltage across the external plate impedance found as soon as its value is known.

The usual method of using a heater type of tube such as the C-327 for grid-circuit detection is to return the grid to the

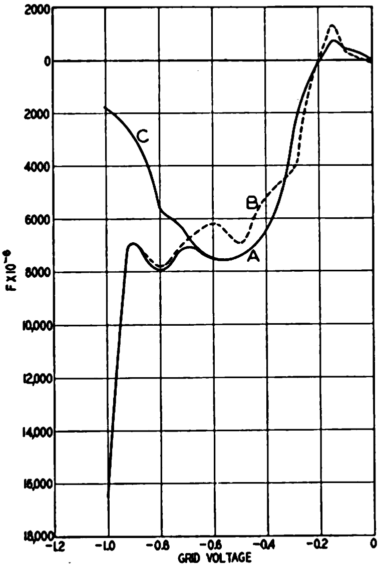

Fig. 5

$$F = \left[- \frac{R_1}{\sqrt{R_1^2\omega^2C_1^2 + 1 + R_1K_g(2 + R_1K_g)}} g_m \frac{\partial K_g}{\partial e_g} + \frac{\partial g_m}{\partial e_g} \right.$$

$C_1 = 250$ $\mu\mu$f.

A—Dynamic Value of *F* Zero Frequency
B—Static Value of *F* Zero Frequency
C—Dynamic Value of *F* 3000 cycles

cathode which is at zero potential. The grid may be returned to any other part of the *B* or *C* supply. It might be of advantage in this type of tube to return the grid to some point of the *B* supply which is at a small positive potential above the cathode. A larger value of grid leak could then be used to pass through some desirable operating point than if the grid is returned to

[7] Stuart Ballantine, "Detection by Grid Rectification," Proc. I. R. E., 16, 593; May, 1928.

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the cathode. In this investigation it was decided to return the grid to the cathode and try various values of R_1.

Fig. 5A and 5B shows the part of (7) in parenthesis plotted for different values of R_1 returned to the cathode. The curves were plotted by finding from Fig. 1 the required values of R_1 to pass through, zero and different values of E_c. The value of K_g was then found from Fig. 2. A and B were plotted for zero frequency. The presence of the grid condenser causes frequency distortion. Fig. 5C shows curve A when the audio frequency is 3000 cycles and C is 250$\mu\mu$f. There is quite serious frequency distortion if R_1 is chosen so that it intersects I_g in Fig. 1 at more than 0.7 volt negative.

The mutual conductance for this tube is shown in Fig. 6 and the values of $\partial g_m/\partial e_g$, the main term in plate rectification, are shown plotted in Fig. 7A and 7B. Their effect in changing (Det. E)$_l$ is small over most of the range. The agreement is

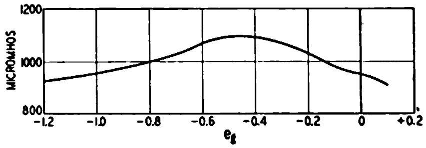

Fig. 6—Mutual Conductance, g_m, of Typical C-327 Tube.
Eb—45 v.
Ef—2.5 v.

not very good between the static and dynamic values. The plate current would change several microamperes making it difficult to obtain the true change in direct current when a signal was applied to the grid.

The dropping off of the mutual conductance as shown by Fig. 6 occurs at a small negative potential. In the usual type of tube the mutual conductance curve straightens out at about 2.5 volts positive, or near the potential of the filament center.

The frequency distortion of this tube used in a popular radio set was checked experimentally for two values of grid leaks 2.0 megohms, the grid leak furnished with the set, and 0.5-megohm grid leak. The set uses a push-pull stage for the last audio stage. The frequency distortion was determined experimentally by the method shown by Ballantine.[7] Briefly this consists of determining the audio amplification from the grid of the detector to the grid of the last tube with the plate

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load or speaker in place. The input is then placed in series with the grid leak with the low side grounded. Because of the shape of the mutual conductance curve it was found to be necessary to modify Ballantine's method slightly. Instead of shorting the grid leak R to determine the audio amplification it was bypassed with a large condenser. This keeps the same bias on the tube for both measurements. This is important if the mutual conductance changes much from the operating

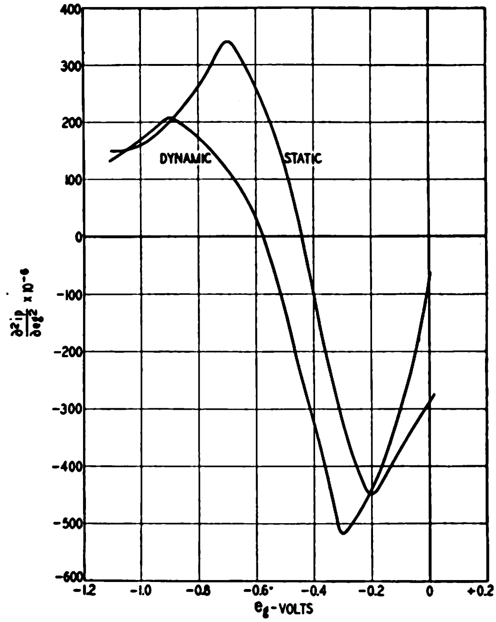

Fig. 7—Variation of Mutual Conductance, $\partial g_m/\partial e_g$, of Typical C-327 Tube.
Eb—45 v.
Ef—2.5 v.
A—Static Value
B—Dynamic Value

point to zero bias. After the two curves were determined experimentally the factor to bring them together at a low frequency, in this case 60 cycles, was determined. The values of the curve determined by the input being in series with the grid leak were then multiplied by this factor. The difference between the values at the higher frequency represents the loss caused by frequency distortion of the detector.

Curve A of Fig. 8 shows the audio amplification from the grid of the detector up to the grid of one tube of the push-pull

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stage determined as shown above. Fig. 8B shows the curve obtained by placing the input in series with the grid leak after the two curves are brought together at 60 cycles. The grid leak used in this case is the 2.0-megohm grid leak furnished with the set. Fig. 8C shows the same curve after a ½-megohm grid leak was substituted for the 2.0-megohm grid leak. The loss in amplification due to the frequency distortion of the detector is somewhat greater than that shown by *C* and *B* of Fig. 5 at 3000

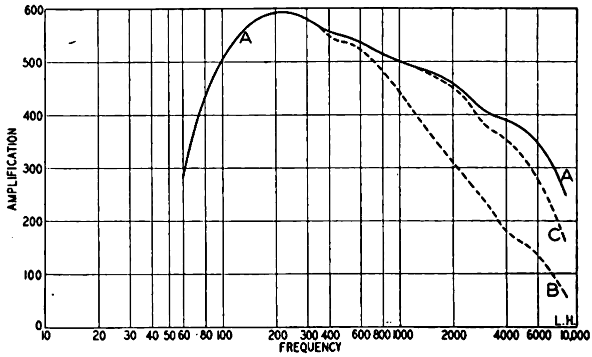

Fig. 8—Audio Amplification and Detector Frequency Distortion of a Commercial Set.
A—Amplification grid of detector to grid second audio tube.
B—Detector frequency distortion with 2.0-megohm grid leak.
C—Detector frequency distortion with 0.5-megohm grid leak.

cycles. This difference is caused by the input capacity of the tube at audio frequencies, which was neglected in calculating *B* of Fig. 5.

The method of determining the values of $\partial K_g/\partial e_g$ at different grid voltages by the change in direct current offers a simple and accurate means of evaluating this factor. The results bear out the theory for small input voltages. No attempt was made to find the range of input voltages over which the theory would hold.

The value of grid leak should be chosen to fit the type of tube used and the values of d.c. voltages used on it. *A* and *C* of Fig. 5 show graphically the range of grid voltage over which the detector will be most sensitive as far as grid detection is concerned, considering both low and high frequency. For example, according to these curves the detector sensitivity would be about the same from −0.4 to −0.7 volt; above −0.7 volt detector frequency distortion becomes serious.

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There is another factor to consider also. The conductance at -0.4 volt is 130×10^{-6} mhos and at -0.7 volt it is about 10×10^{-6} mhos. The value of -0.7 volt would be preferable to use as the selectivity of the circuit preceding the detector would be very poor with 130×10^{-6} mhos shunted across it. The value of grid leak to use may be found from the static characteristic of Fig. 1 as soon as the operating point is found from Fig. 5.

The experimentally determined curves of Fig. 8 bear out the theoretical considerations that the grid leak should be chosen to match the tube characteristics. The grid leak should be as high a value as possible considering only the radio-frequency tuned circuit preceding the detector. The value of the grid leak should be as low as possible considering frequency distortion of the detector. Some value in between will be the best compromise. In the tube analyzed a value of resistance that would bias the tube at about -0.7 volt would be the best to use.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

THE RADIATION RESISTANCE OF BEAM ANTENNAS*

By
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Summary—*In this paper a new method proposed by Brillouin for the calculation of radiation resistance is applied to several types of beam antennas. New formulas are deduced and some interesting results are obtained showing the distribution of the radiated power among the different wires of beam antennas and giving the numerical value of the radiation resistance in various cases (synphase antenna, antiphase antenna, Marconi three-stage antenna). The radiation resistance in the presence of a perfect conducting plane is also considered. A table of values of the components of radiation resistance is added to the paper for practical use.*

THERE are two methods of computing the power radiated by an antenna. In the first we calculate the Poynting-vector for each point of space and integrate the normal energy flow through any surface enclosing the antenna. This method might be called the Poynting-vector method, and has been used in the well known works of G. W. Pierce,[1] B. van der Pol,[2] S. Ballantine,[3] M. A. Bontsch-Bruewitsch,[4] and S. Levin and C. Young.[5] It may be noticed that we cannot by this method obtain the contributions to the radiated power of different parts of the antenna, which it is sometimes desirable to know when dealing with some practical cases.

The other method is based upon the study of the emfs induced in the antenna by the currents in the wires of which the antenna is constructed. Let AB (Fig. 1) be a wire in which flows a current of frequency f and assume that the distributions of the current and of the charges in the wire are known. We may then calculate the electric force in each point M of the space. It depends upon the combined action at this point of all elements of the wire. In particular we may take the point M_0 to lie in the surface of the wire itself and calculate the emf due to the electromagnetic field of the wire. If we assume the action at a distance of the current to be instantaneous, this emf would

* Dewey decimal classification: R125.6. Original manuscript received by the Institute, May 25, 1928. Revised manuscript received, October 8, 1928.

[1] G. W. Pierce, *Proc.* Amer. Acad. **52**, 192, 1916.
[2] Balth. van der Pol, Jr., *Proc.* Phys. Soc. of London, **13**, **217**, 1917.
[3] Stuart Ballantine, Proc. I. R. E., **12**, 823, 1924; **15**, 245, 1927.
[4] M. A. Bontsch-Bruewitsch, *Annalen der Physik*, **81**, 425, 1926.
[5] S. A. Levin and C. J. Young, Proc. I. R. E., **14**, 675, 1926.

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be a purely reactive one and we may speak of it as arising from the capacity or the inductance of the wire. If the wire is of a length comparable with that of the electromagnetic wave we must take into account the propagation velocity of the field. Then at a moment t there will act in the point M the current and the charge which have existed at the element dx_1 at the earlier moment $\left(t-\frac{r_1}{c}\right)$, r_1 being the distance of the point from dx_1, and c the light-velocity. Similarly for the element dx_2 we must use the values of current and charge existing at the earlier moment $\left(t-\frac{r_2}{c}\right)$. Under these conditions the calculated emf (or, strictly speaking, the corresponding potential drop) will have a watt component which may be called the radiation emf.

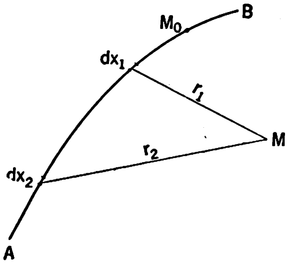

Fig. 1

The product of this emf and the current in an element of the wire gives the radiation from the element. We may find by integration the expenditure of power for radiation in the whole antenna or in its different parts.

We shall call this method the induced emf method. It was proposed by Brillouin[6] and applied by Kliatzkin[7] in the analysis of the radiation of a vertical earthed wire. It is based upon the electromagnetic field equations in the form employing the retarded potentials of Lorentz.[8]

This paper deals with the radiation resistance of antennas, composed of parallel half-wave vibrators.

[6] *Radioélectricité*, April, 1922.
[7] *Telegrafia i telefonia bez provodov* (TITBP) 1(40), 33, 1927.
[8] Lorentz, "The Theory of Electrons," 2 ed., Chap. 1, paragraphs 13 and 14.

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1. Outline of the Method. Radiation EMF. We have first to solve the following problem: a single wire with a known distribution of current being specified, a formula is to be found for the component of electric force parallel to the wire at any point of space.

This may be derived from the expression given by Lorentz:

$$E=-\frac{1}{c}\frac{\partial A}{\partial t}-\operatorname{grad}\phi\,; \tag{1}$$

where

$$\phi=\frac{1}{4\pi}\int\frac{1}{r}[\rho]dv \tag{2}$$

is a scalar potential at the given point due to the fixed charges distributed in the space and

$$A=\frac{1}{4\pi c}\int\frac{1}{r}[\rho v]dv \tag{3}$$

is a vector potential due to the charges, moving with the velocity v.

Applying this to the case of a very thin straight wire and passing from the Lorentz units to the absolute system of units we obtain

$$\phi=\frac{1}{\epsilon}\int_0^l\frac{|\sigma|_{t-r/c}}{r}dx \tag{4}$$

$$A=\mu c\int_0^l\frac{|i|_{t-r/c}}{r}dx \tag{5}$$

where σ is the charge per unit length, i is the current in the element dx of the wire, l is the length of the wire, and r is the distance of the given point from dx. The values of charge and current must be taken at the time $\left(t-\dfrac{r}{c}\right)$.

Let the origin of coordinates lie at the one end of the wire and let the OX axis lie along the wire (Fig. 2). The component of the electric force parallel to the wire will then be at the point M (∂, ξ) as follows:

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$$E=-\frac{1}{c}\frac{\partial A}{\partial t}-\frac{d\phi}{d\xi}=-\int_0^l\frac{\mu}{r}\frac{\partial i_{(t-r/c)}}{\partial t}dx-\frac{\partial}{\partial\xi}\int_0^l\frac{\sigma_{(t-r/c)}}{\epsilon r}dx \qquad (6)$$

If the angular frequency is ω then $i=I_x\sin\omega t$ and

$$\frac{\partial i(t-r/c)}{\partial t}=\omega I_x\cos\omega\left(t-\frac{r}{c}\right) \qquad (7)$$

$$\sigma_{(t-r/c)}=-\int\frac{\partial i(t-r/c)}{\partial x}dt=\frac{1}{\omega}\frac{\partial I_x}{\partial x}\cos\omega\left(t-\frac{r}{c}\right) \qquad (8)$$

from the equation $-\frac{\partial\sigma}{\partial t}=\frac{\partial i}{\partial x}$ giving the relationship between the

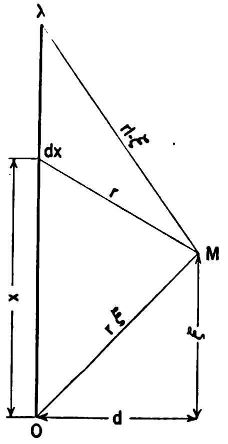

Fig. 2

current and the charge along the wire.

The instantaneous value of electric force will be as follows

$$e_d=-\mu c\left[m^2\int_0^l\frac{\cos(\omega t-mr)}{mr}I_xdx+\frac{\partial}{\partial\xi}\int_0^l\frac{\cos(\omega t-mr)}{mr}\frac{\partial I_x}{\partial x}dx\right] \qquad (9)$$

where $m=\frac{\omega}{c}=\frac{2\pi}{\lambda}$ and $r=\sqrt{d^2+(x-\xi)^2}$.

Assume that the current is distributed sinusoidally along the wire and is zero at the origin of coordinates. Thus

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$$I_x = I_0 \sin mx \tag{10}$$

where I_0 is the amplitude of the current at the loop.

The expression for e_d will then be

$$e_d = -\mu c\left[m^2 \int_0^l \frac{\cos(\omega t - mr)}{mr} I_0 \sin mx dx + m\frac{\partial}{\partial \xi}\int_0^l \frac{\cos(\omega t - mr)}{mr} I_0 \cos mx\, dx\right] \tag{11}$$

After integration we obtain:

$$e_d = \mu c I_0\left[\frac{\cos(\omega t - mr_{l-\xi})}{r_{l-\xi}} \cos ml - \frac{\cos(\omega t - mr_\xi)}{r_\xi}\right] \tag{12}$$

where $r_{l-\xi} = \sqrt{d^2 + (l-\xi)^2}$ and $r_\xi = \sqrt{d^2 + \xi^2}$.

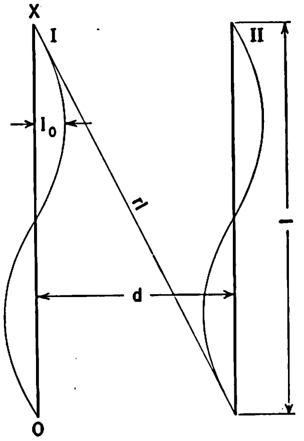

Fig. 3

In the particular case in which the point M lies on the wire itself, $d = 0$ and we shall have:

$$e_0 = \mu c I_0\left[\frac{\cos(\omega t - ml + m\xi)}{l - \xi} \cos ml - \frac{\cos(\omega t - m\xi)}{\xi}\right]. \tag{13}$$

2. Case of Two Parallel Wires. We shall now study the problem of the radiation of power from a system formed by two parallel wires. For this purpose we may assume some conditions which simplify the solution. We shall consider (Fig. 3) equal

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wires whose lengths are multiples of the half-wavelength. The wires are not displaced in height, that is, their ends must lie on the straight line perpendicular to the direction of the wires. The distribution, the phase and the values of current we assume to be identical in both wires.

Under these conditions the tangential component of E at any point along either wire comprises two parts, the first produced in each one by its own current and a second produced by the current in the other wire: i.e., $e=e_0+e_d$. This electric force produces an emf in the wire, and the power needed to suppress it will be the radiation power. For the element dx of the wire this power will be

$$dP_\Sigma = -EI_x \cos\phi dx \tag{14}$$

where E and I_x are the effective values of the electric force and current and ϕ is the phase angle.

The total power for one wire having the length l will be

$$\begin{aligned} P_\Sigma = -\int_0^l EI_x \cos\phi dx = -\int_0^l E_0 I_x \cos\phi_0 dx \\ -\int_0^l E_d I_x \cos\phi_d dx = P_0 + P_d. \end{aligned} \tag{15}$$

We shall first calculate P_d. Let the full expression of E and I_x be written. To obtain the power in watts we must take the current in amperes and E_d, E_0 in volts. Then

$$I_x = I_0 \sin mx \tag{16}$$

where I_0 is the effective value of the current at the loop in amperes.

E_d is also the rms value and from the formula (12) we obtain two components of it

$$E' = 30I_0 \frac{\cos ml}{r_{l-x}} \quad \text{and} \quad E'' = 30I_0 \frac{1}{r_x} \tag{17}$$

Each of these components has a different phase angle ϕ_1 and ϕ_2. Let us find them.

From the expression (12) we have

$$e_d' = E_d' \cos(\omega t - mr_{l-x}) = E_d' \sin\left(\omega t - mr_{l-x} + \frac{\pi}{2}\right). \tag{18}$$

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The current in the wire is $i_x = I_x \sin \omega t$

Therefore $\phi_1 = mr_{l-x} - \frac{\pi}{2}$ and

$$\cos \phi_1 = \cos\left(mr_{l-x} - \frac{\pi}{2}\right) = \sin mr_{l-x} \tag{19}$$

Similarly

$$\cos \phi_2 = \sin mr_x.$$

Thus we obtain for P_d

$$P_d = -30I_0^2 \int_0^l \left(\frac{\sin mr_{l-x}}{r_{l-x}} \cos ml - \frac{\sin mr_x}{r_x}\right) \sin mx\, dx \tag{20}$$

By integrating we obtain the following expression for the radiated power *when the length of the wire is a multiple of the half-wavelength.*

$$P_d = 30I_0^2[2Ci\, md - Ci\, m(\sqrt{d^2+l^2}+l)$$
$$-Ci\, m(\sqrt{d^2+l^2}-l)] = 30I_0^2 M_d \tag{21}$$

Here $Ci(x)$ denotes the integral cosine, d is the distance between the wires, l is the length of the wire.

P_d is merely one of the components of radiation power, depending upon the current in the other wire. The second component P_0 we may obtain as limit of P_d when the distance between the wires approaches O.

$$P_0 = \lim |P_d| = 30I_0^2(E + \log 2ml - Ci2ml) = 30I_0^2 M_0 \tag{22}$$

where $E = 0.577 \cdots$ is the Eulers constant.

The whole radiation power in one wire will be

$$P = P_0 + P_d \tag{23}$$

and the radiation power of the system of two wires

$$P_\Sigma = 2P = 2P_0 + 2P_d \tag{24}$$

Dividing the expressions (21) and (22) by I_0^2 gives the so-called "radiation resistance." Obviously we may speak of this quantity only when the currents in both wires are equal.

3. Application to Beam Antennas. Synphase System. We shall now apply these results to the computation of the radiation resistance for some types of beam antennas. We shall consider

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first the so-called synphase antenna, composed of single parallel vibrators (Fig. 4). The vibrators are situated along a straight line at a distance of a half-wavelength from each other. Their currents are equal and in phase. In this case the beam has a direction perpendicular to the line of the wires.

The power radiated by any individual antenna wire is composed of the power due to its own current and that due to the electric force induced in it by other vibrators. We shall denote

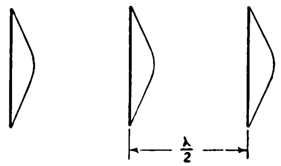

Fig. 4

by P_d the power corresponding to induction by a vibrator at a distance d; in the case of the antenna considered d will be a multiple of a half-wavelength. As stated above in (21),

$$P_d = 30 I_0^2 M_d$$

As the currents in all wires are equal the component of radiation resistance due to another wire will be

$$R_d = 30 M_d.$$

The values of these components are given in Table A (line 1) for distances d which are multiples of the half-wavelength.[9]

Using this table let us now compute the radiation resistance for the antenna composed of three vibrators.

For each of the extreme wires we shall have:

$$R_1 = R_3 = R_0 + R_{\lambda/2} + R_\lambda = 73.3 - 12.4 + 4.1 = 65.0\ \Omega$$

and for the middle wire:

$$R_2 = R_0 + 2R_{\lambda/2} = 73.3 - 2 \times 12.4 = 48.5\ \Omega.$$

The radiation resistance of the whole antenna will be:

$$R_3 = 3R_0 + 4R_{\lambda/2} + 2R_\lambda = 178.5\ \Omega.$$

[9] For these calculations I employed the integral-function curves especially plotted by Mr. E. D. Milovidoff of the staff of the Nijni-Novgorod Radiolaboratory. Interpolation from commonly used tables (Jahnke u. Emde) is rather misleading.

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For the four-wire antenna we obtain by analogous calculations:

$$R_2 = 4R_0 + 6R_{\lambda/2} + 4R_\lambda + 2R_{3\lambda/2}$$

and generally for an antenna composed of n wires:

$$R_n = nR_0 + 2(n-1)R_{\lambda/2} + 2(n-2)R_{2\lambda/2} + \cdots + 2R_{(n-1)\lambda/2} \quad (25)$$

Table I contains values of the radiation resistance (a) for each vibrator, (b) for the whole antenna, and (c) the mean value for one vibrator. It might be noticed that when increasing the number of vibrators the last quantity is very rapidly approaching the limit (near 56 ohms), which was obtained for the case of an infinitely great number of vibrators by using the "Poynting vector method" of radiation resistance calculation.[10]

TABLE I

Values of radiation resistance in ohms for synphase beam antenna. n = number of wires; R_n = resistance of nth wire; R = total resistance. R_m = average resistance per wire.

n	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R	R_m
2	60.9	60.9	—	—	—	—	—	121.8	60.9
3	65.0	48.5	65.0	—	—	—	—	178.5	59.5
4	63.2	52.6	52.6	63.2	—	—	—	231.8	58.0
5	64.4	50.9	56.7	50.9	64.4	—	—	287.3	57.4
6	63.6	52.0	55.0	55.0	52.0	63.6	—	341.4	56.9
7	64.0	51.2	56.1	53.2	56.1	51.2	64.0	395.6	56.5

4. Continuation. Antiphase Beam Antenna. We shall now pass on to the antiphase antenna (Fig. 5) which radiates a beam directed in the plane of antenna. It differs from the synphase

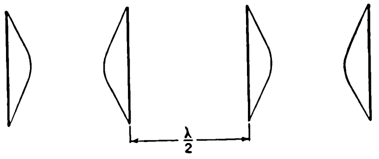

Fig. 5

antenna only in the fact that the currents in the adjacent wires have a phase difference of 180 deg. All the radiation resistance components having d equal to an odd number times the half-wavelength must therefore be multiplied by -1. The other components are the same as before because vibrators spaced by an integral number of wavelengths are in phase.

[10] M. A. Bontsch-Bruewitsch, l. c., p. 434.

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The radiation resistance for this antenna is expressed by the following general formula:

$$R_n = nR_0 - 2(n-1)R_{\lambda/2} + 2(n-2)R_{\lambda} - \cdots + 2R_{(n-1)\lambda/2} \quad (26)$$

where the quantities R_d may be taken from Table A (line 1). The results of calculations for the antiphase antenna are given in Table II.

From Tables I and II we may see that the radiation resistance is different for the various wires in the antenna and this difference is unequal for the two types of antennas. As the number of vibrators is increased the difference diminishes.

TABLE II

Value of radiation resistance in ohms for antiphase beam antenna (for explanation see legend of Table I).

n	R_1	R_2	R_3	R_4	R_5	R_6	R_7	R	R_m
2	85.7	85.7	—	—	—	—	—	171.4	85.7
3	89.8	98.1	89.8	—	—	—	—	277.6	92.5
4	91.5	102.2	102.2	91.5	—	—	—	387.2	96.8
5	92.7	103.9	106.2	103.9	92.7	—	—	499.2	99.8
6	93.4	105.1	108.0	108.0	105.1	93.4	—	612.8	102.1
7	93.9	105.8	109.1	109.7	109.1	105.8	93.9	727.2	103.9

5. Parallel Wires Displaced in Height. As a next step in the development of this method, the radiation resistance of parallel vibrators displaced in height may be calculated, (Fig. 6). The investigation of this case will enable us to study the radiation resistance in the presence of a perfectly conducting plane and to calculate the radiation resistance of multistage antennas.

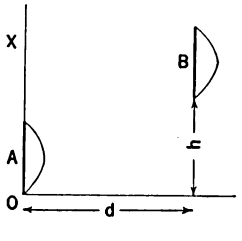

Fig. 6

We will deduce a formula for the radiation power due to the emf induced in the vibrator B by another vibrator A. The wire B is placed a distance d and elevated on a height h with respect to A. We shall denote this power by $P(d, h)$ and the corresponding radiation resistance component by $R(d, h)$.

The electric force near B due to the first vibrator is defined by (12).

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The law of current distribution in the wire B is as follows:

$$I_x = I_0 \sin m(x-h) \tag{27}$$

Proceeding exactly as in Sect. 2 we shall obtain for the radiation power due to the induction the following expression:

$$P(d,h) = 30 I_0^2 \int_h^{h+\lambda/2} \left(\frac{\sin mr_{l-x}}{r_{l-x}} + \frac{\sin mr_x}{r_x} \right) \sin m(x-h)dx. \tag{28}$$

The integration gives for the corresponding radiation resistance component a rather complicated expression, as follows:

$$\begin{aligned} R(d,\, h) = &-15 \sin mh \cdot \left[S\left(d, h - \frac{\lambda}{2}\right) - 2S(d,\, h) \right. \\ &\left. + S\left(d, h + \frac{\lambda}{2}\right) \right] \\ &-15 \cos mh \cdot \left[C\left(d, h - \frac{\lambda}{2}\right) - 2C(d,\, h) \right. \\ &\left. + C\left(d, h + \frac{\lambda}{2}\right) \right] \end{aligned} \tag{29}$$

where $S(x,\, y)$ and $C(x,\, y)$ are the functions:

$$\begin{aligned} S(x,\, y) &= Si\, m(\sqrt{x^2+y^2}+y) - Si\, m(\sqrt{x^2+y^2}-y) \\ C(x,\, y) &= Ci\, m(\sqrt{x^2+y^2}+y) + Ci\, m(\sqrt{x^2+y^2}-y). \end{aligned} \tag{30}$$

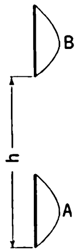

Fig. 7

This formula is the most general one for the case of two parallel vibrators. The expression (21), obtained for the vibrators placed at the same height, is a particular case of it for $h=0$. For the other particular case, when $d=0$ (Fig. 7) we find:

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$$R(0, h) = -15 \sin mh \left[Si\,2m\left(h-\frac{\lambda}{2}\right) - 2Si\,2mh + Si\,2m\left(h+\frac{\lambda}{2}\right)\right] - 15 \cos mh \left[\log\frac{h^2}{h^2-\frac{\lambda^2}{4}} + Ci\,2m\left(h-\frac{\lambda}{2}\right) - 2Ci\,2mh + Ci\,2m\left(h+\frac{\lambda}{2}\right)\right] \tag{31}$$

which is in agreement with the analogous formula obtained by M. A. Bontsch-Bruewitsch.

Having any given complex antenna system comprised of n synphase parallel vibrators we can by means of expressions (29) calculate the radiation resistance for every one of them. This resistance will be:

$$R_\Sigma = R(0,\ 0) + R(d_1\ h_1) + R(d_2,\ h_2) + \cdots + R(d_{n-1},\ h_{n-1}) \tag{32}$$

where h and d denote the height difference and the distance between the first vibrator and each other one; $R(O,O) = 73.3\Omega$.

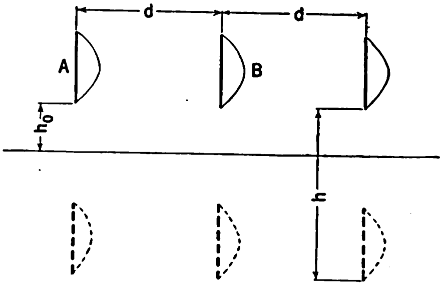

Fig. 8

6. Antenna Over Perfectly Conducting Plane. The expression (29) may also be used for the calculation of the radiation resistance of an antenna erected over a perfectly conducting plane by application of the simple image theory. We shall treat the case of an antenna of which the vibrators are placed at the same height over the plane and at equal distances d from each other.

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Let h_0 be the height of antenna over the conducting plane (Fig. 8). Introducing the correction due to the images we shall obtain:

For each of the extreme wires:

$$R_{\Sigma A}=73.3+R(0,h)+R(d,0)+R(d,h)+R(2d,0)$$
$$+R(2d,h)+\cdots+R[(n-1)d,0]+R[(n-1)d,h] \quad (33)$$

where $h=2h_0+\frac{\lambda}{2}\cdot$

For each wire second from the edge

$$R_{\Sigma B}=73.3+R(0,h)+2[R(d,0)+R(d,h)]$$
$$+\cdots+R[(n-2)d,0]+R[(n-2)d,h]. \quad (34)$$

If the radiation resistance of the whole antenna is to be found we can use a formula analogous to the formula (25)

$$R_\Sigma=nR_0+2(n-1)R_1+2(n-2)R_2+\cdots+2R_{n-1} \quad (35)$$

where $$R_k=R\left(\frac{k\lambda}{2},\ 0\right)+R\left(\frac{k\lambda}{2},\ 2h_0+\frac{\lambda}{2}\right).$$

The calculations were carried through by the author for a synphase 7-wire antenna elevated 0, $\frac{\lambda}{8}$, $\frac{\lambda}{4}$, $\frac{3\lambda}{8}$, and $\frac{\lambda}{2}$ over the plane. The results are shown in Table III. We may notice that with increasing height the total radiation resistance rapidly

TABLE III

h_0	$r_1=r_7$	$r_2=r_6$	$r_3=r_5$	r_4	R	r_m
0	84.7	58.2	74.7	62.0	497.3	71.0
$\frac{\lambda}{8}$	65.0	41.8	59.8	43.7	376.8	53.8
$\frac{\lambda}{4}$	62.8	44.7	58.8	42.2	374.8	53.5
$\frac{3\lambda}{8}$	65.8	50.5	59.0	51.2	401.8	57.4
$\frac{\lambda}{2}$	66.4	51.2	55.5	54.4	400.6	57.2
∞	64.0	51.2	56.1	53.2	395.6	56.5

approaches the value obtained for free space, but the energy distribution between the individual wires is different. As expected the radiation resistance increases near the plane.

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7. Multistage Antenna. The method of induced emfs may be applied to more complicated systems, particularly multistage antennas. Let us consider for example a three-stage antenna of the type employed by Marconi.

A unit of such an antenna is a system of three synphase vibrators, spaced along a straight line. These vibrators are connected through antiresonant coils. Let us compute the power due to the emf induced by such an antenna unit in another one spaced at a distance d (Fig. 9). We assume the currents to be equal and in phase.

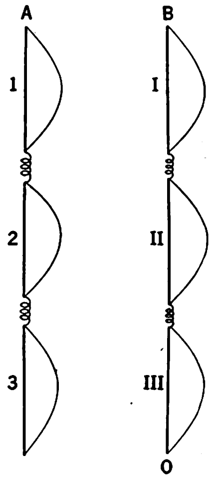

Fig. 9

The power induced in vibrator I of the wire B may be resolved into three parts due to the vibrators 1—3, respectively, of the wire A. Using our notation we may write:

$$P_{AI} = P(d,0) + P\left(d, \frac{\lambda}{2}\right) + (P,\lambda) \tag{36}$$

Similarly for the II and III vibrators:

$$P_{AII} = P(d,0) + 2P\left(d, \frac{\lambda}{2}\right)$$
$$P_{III} = P_{AI}. \tag{37}$$

The whole power induced in the wire B will be

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$$P_{AB}=3P(d,\lambda)+4P\left(d,\frac{\lambda}{2}\right)+2P(d,\lambda) \tag{38}$$

and generally in the case of n-stage wire

$$P_{AB}=nP(d,0)+2(n-1)P\left(d,\frac{\lambda}{2}\right)+2(n-2)P(d,\lambda)$$
$$+\cdots+2P\left(d,(n-1)\frac{\lambda}{2}\right). \tag{39}$$

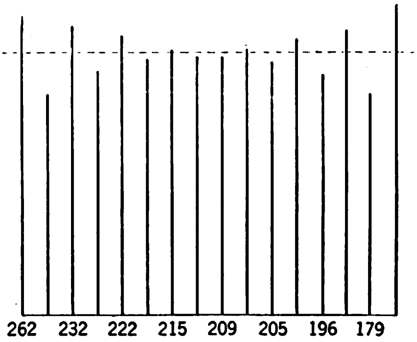

Fig. 10

The radiation resistance of single three-stage wire is obtained by substituting $d=0$.

$$P_0=3P(0,0)+4P\left(0,\frac{\lambda}{2}\right)+2P(0,\lambda)=317.1\,\Omega \tag{40}$$

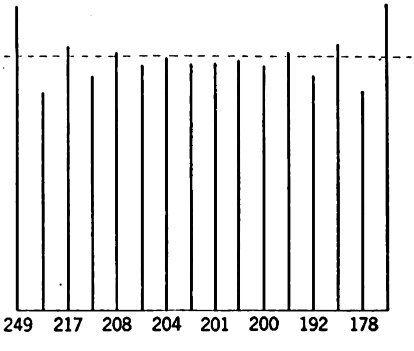

Fig. 11

In order to calculate the radiation resistance of different wires and of the whole antenna, formulas may be used analogous to those obtained above.

The author has performed these calculations for the case of a 16-wire antenna, the distance between the wires being assumed to be $\frac{\lambda}{2}\cdot$ The results are shown graphically in Fig. 10; the numbers below give the radiation resistance in ohms. The mean value of this resistance (214 ohms) is marked by a dotted line. Analogous calculations were also made for this antenna elevated at $\frac{\lambda}{4}$ over a perfectly conducting earth. The mean value of the radiation resistance for one wire is then 206 Ω. The energy distribution is shown in Fig. 11. In both cases this distribution

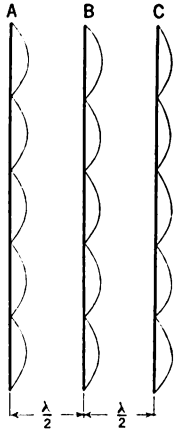

Fig. 12

is very nonuniform in the extreme wires. This means that the design of the feeding devices for the several wires must be quite different, if we wish to obtain equal currents in all vibrators.

8. Table A. Various other types of antennas may be computed in the same manner. To simplify the calculations a table is appended (Table A), containing the functions $R(d, h)$ for values of d and h, which are multiples of a half-wavelength. This table will be found useful in the calculation of many practical types of directive antennas.

As an example let us calculate the radiation resistance of an antenna formed by three five-stage wires (Fig. 12) spaced at a

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distance of a half-wavelength from each other. Proceeding exactly as in Sect. 7 for the three-stage antenna we shall obtain for the radiation resistance component R_1 of the wire A due to the wire B the following expression:

$$R_1 = 5R\left(\frac{\lambda}{2}, 0\right) + 8R\left(\frac{\lambda}{2}, \frac{\lambda}{2}\right) + 6R\left(\frac{\lambda}{2}, \lambda\right) + 4R\left(\frac{\lambda}{2}, \frac{3}{2}\lambda\right) + 2R\left(\frac{\lambda}{2}, 2\lambda\right) \quad (41)$$

The values of $R(d, h)$ may be taken from the table. Thus we obtain

$$R_1 = 5\cdot(-12.36) + 8\cdot(-11.80) + 6\cdot(-0.78) + 4\cdot(+0.80) + 2\cdot(-1.00) = -159.7\,\Omega \quad (42)$$

TABLE A

$h \backslash d$	0	0.5	1.0	1.5	2.0	2.5	3.0	3.5
0	+73.29	−12.36	+4.08	−1.77	+1.18	−0.75	+0.42	−0.33
0.5	+26.40	−11.80	+8.83	−5.75	+3.76	−2.79	+1.86	−1.54
1.0	− 4.065	− 0.78	+3.56	−6.26	+6.05	−5.67	+4.51	−3.94
1.5	+ 1.78	+ 0.80	−2.92	+1.96	+0.16	−2.40	+3.24	−3.76
2.0	− 0.96	− 1.00	+1.13	+0.56	−2.55	+2.74	−2.07	+0.74
2.5	+ 0.58	+ 0.45	−0.42	−0.96	+1.59	−0.28	−1.59	+2.66
3.0	− 0.43	− 0.30	+0.13	+0.85	−0.45	−0.10	+1.74	−1.03

TABLE A (Cont'd)

$h \backslash d$	4.0	4.5	5.0	5.5	6.0	6.5	7.0	7.5
0	+0.21	−0.18	+0.15	−0.12	+0.12	−0.10	+0.06	−0.03
0.5	+1.08	−0.85	+0.69	−0.57	+0.51	−0.45	+0.36	−0.30
1.0	+3.08	−2.50	+2.10	−1.80	+1.56	−1.18	+1.14	−1.00
1.5	+3.68	−3.40	+3.14	−2.90	+2.61	−2.31	+2.06	−1.86
2.0	+0.51	−1.30	+1.82	−2.24	+2.28	−2.29	+2.26	−2.14
2.5	−2.49	+2.00	−1.35	+0.49	−0.06	−0.45	+0.85	−1.03
3.0	−0.09	+1.12	−1.87	+1.77	−2.02	+1.71	−1.32	+0.66

The values of d and h are given here in parts of the wavelength; those of radiation resistance in ohms.

To find the radiation resistance component R_2 due to the wire C or the radiation resistance of the single five-stage wire R_0 we must substitute in the above expression $d=\lambda$ or $d=0$ instead $d=\frac{\lambda}{2}\cdot$ Therefore

$$R_2 = +103.0\,\Omega \text{ and } R_0 = 558.5\,\Omega$$

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The radiation resistance of the wires A and C will then be:

$$R_A = R_C = R_0 + R_1 + R_2 = 501.8\ \Omega$$

and of the wire B

$$R_B = R_0 + 2R_1 = 239.1\ \Omega\ .$$

The whole radiation resistance of the antenna will be 1242.7 Ω, the mean value for 1 wire = 416.2 Ω and for 1 vibrator = 83.0 Ω.

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Discussion on
SIMPLE INDUCTANCE FORMULAS FOR RADIO COILS*

(Harold A. Wheeler)

R. R. Batcher[1]: A formula shown in this paper* for the rapid computation of the inductance of air-core radio coils corresponds to one which the writer has used for about six years. Having found by experience that many engineers to whom this and other simple formulas have been disclosed have an aversion to using them because they have insufficient accuracy, it is believed that further analysis might be of interest.

The formula

$$L = an^2Q \tag{1}$$

has been called the universal inductance formula since tables, formulas, and charts have been derived to cover almost every shape of air-core inductance in evaluating Q, which is a factor depending entirely on the physical dimensions of the coil. For a single layer solenoid the Lorenz formula may be used or the tables which have been obtained from it for representative values. This formula is one of the most exact in its field.

An inspection of a curve for Q plotted from these tables indicates that it is probably of a hyperbolic nature. If so a comparison of the relation between $1/Q$ and $b/2a$ would be a straight line. Analysis of this line gives an empirical value for Q (after converting to inch measurements by multiplying by 2.54).

$$Q = \frac{100}{(b/a+0.9)} \tag{2}$$

Substituting (2) in (1) gives

$$L = \frac{a^2n^2 1000}{10b+9a} \text{ centimeters} \tag{3}$$

$$= \frac{a^2n^2}{10b+9a} \text{ microhenries} \tag{4}$$

which is the formula previously reported. The deviation between the curve of $1/Q$ and a straight line is a measure of the accuracy of this equation. For greater accuracy with very short coils the denominator may be changed to $(8.3a+10b)$.

It may be shown that the form of (4) could be obtained from fundamental considerations, and since it brings to light interesting facts it will be derived. One of the basic inductance relations is

$$\text{Inductance} = \frac{4\pi n^2}{\text{Reluctance of path}} \tag{5}$$

For an air core with unit permeability the reluctance is the value of (length of path/area of path). It is this relation that makes the ordinary

* Proc. I. R. E., 16, 1398; October, 1928.
[1] Decatur Mfg. Co., Brooklyn, N. Y.

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inductance formula so complicated since both the length and area of the flux path are indeterminate. But the *effective* length and the *effective* area values are interrelated so that their quotient is a constant for any one coil. It is convenient, therefore, to assume a value for the area and from this determine the length of the flux path. It may be assumed that the area of the path is the area enclosed by the average turn, i.e., area $=\pi a^2$, throughout the whole length of the flux path. Then the length of the flux path may be taken to be equal to $(b+q)$ where b is the length of the coil and q is a fictitious length equal to the length of the return path of the flux outside of the coil. Substituting these values in (5) gives

$$L=\frac{4\pi n^2}{\left\{\frac{b+q}{\pi a^2}\right\}}\ 2.54=\frac{(2\pi a)^2 2.54 n^2}{b+q} \tag{6}$$

Comparing (6) with (4) indicates that the flux return path for the values assumed is equal to $0.9a$ (which may appear paradoxical but the value is due to the small area of the flux path assumed exterior to the coil).

This same procedure has been applied to many other types of coils and may be applied to many more. In many cases it is desirable to derive a simple inductance formula for a system of coils which may be under consideration. In one case such a relation was simply obtained for a special shape of coil antennas which assisted materially with the experiments. If a series of measurements are made to cover representative conditions such a formula may be readily derived.

(a) Determine the value of $1/Q$ from (1) using the available data, for several values of $2a/b$.

(b) Determine (k) from two representative values of Q, say Q' and Q'' with their corresponding shape ratios b'/a' and b''/a'', using

$$k=(b'/a' - b''/a'')\frac{Q'Q''}{Q'-Q''} \tag{7}$$

(c) Determine q for several values of Q from

$$q=1/Q-b/ak \tag{8}$$

The value of q should not change very much over a considerable range of values for $b/2a$.

A multilayer coil presents greater difficulties. Several have been disclosed, such as the Brooks-Turner formula, which is probably the one most commonly used. As a first approximation it can be shown that the geometric mean distance from a point within a rectangle to the rectangle is very nearly equal to $0.223\ (b+c)$. Here c is the winding depth. The g.m.d. from a point on a line to the line is equal to $0.223\ b$. From this it would seem that the factor $(b+c)$ could be substituted in the single layer formula for the factor b, to obtain a multilayer formula:

$$L=\frac{a^2n^2}{9a+10(b+c)} \tag{9}$$

This formula gives fair accuracy for short coils, but in some other cases a large error results. A more accurate analysis gives:

$$L=\frac{a^2n^2}{9a+10(b+c)\dfrac{2bc}{a}} \tag{10}$$

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It has been shown[2] that no great error results when a single layer coil inductance formula is used for flat spirals by substituting (c) for (b). Formula (4) may be used.

It should be kept in mind that all of these formulas are based on current sheet considerations and are subject to other corrections. They are thus not true "empirical" formulas, which are ordinarily designed to fit a given system of measurements.

In conclusion it may be well to call attention to the normal accuracy requirements of an inductance formula. I have seen engineers reject an approximate formula with an admitted error of say 3 per cent and reject slide rule computations in favor of long hand computations, but who nevertheless used an ordinary scale graduated in sixteenths to determine the dimensions of the coil. The nominal diameter of the tube was used (3 inches) for $2a$, whereas the actual diameter was later found to be 3.045 in., and adding the diameter of the wire increased the value to 3.095 in. The length was taken as $2\frac{7}{8}$ in., whereas the actual length (nd). was 2.82 in. The total error due to these measurements was over 6 per cent, although the greatest error in the measurements was less than $\frac{1}{16}$ in.

In this light great accuracy in a formula seems in many cases to be of secondary importance.

Harold A. Wheeler[3]: I believe Mr. Batcher gives insufficient attention to two very important details which I carefully noted for both formulas recommended in my brief paper to which he refers. First, what is the maximum error of the approximate formula? and secondly, within what limits of shape factors does this maximum error obtain? Until these questions are answered, no engineer is justified in using an approximate formula.

The type of formula under discussion has no *theoretical* basis for "short" coils $(b+c<2a)$, since only a small part of the total flux links with all the turns. Furthermore, when there are few turns or large spacing between turns, which is often true of "short" coils, one or two corrections cannot be neglected.

Theoretical consideration of a coil of very small winding space yields the following formula of a different type:

$$L = \frac{an^2}{13.5} \log_{10} \left\{ \frac{4.9a}{b+c} \right\} \text{ microhenries} \qquad (4)$$

(dimensions in inches). On the assumption of uniform current distribution in a rectangular cross-section, the error of this formula is less than 3 per cent when $b+c \leqq a$.

Formula (4) is the only type which remains accurate as the term $b+c$ decreases indefinitely. The Hazeltine formula (1) is the most generally accurate of any of its type, as applied to multilayer coils of medium and small winding space.

Since preparing the above paper, I find that in July, 1920, I first derived formula (2) therein, with almost identically the same constants now recommended. The Hazeltine formula (1) therein was derived about 1916 and published shortly thereafter.

[2] J. A. Fleming, "Wireless Telegraphist's Pocketbook," p. 97.
[3] Hazeltine Service Corporation Laboratory, New York City.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 3 *March, 1929*

MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared by the radio laboratory of the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The various articles listed below are not obtainable from the Bureau of Standards. The various periodicals can be consulted at large public libraries.

R100. Radio Principles

R113 Radio transmission and the upper atmosphere (editorial). *Experimental Wireless and Wireless Engineer* (London), 5, 657–659; December, 1928.

(Abstract of a paper by Appleton comparing the three methods for determining the effective height of the Heaviside layer. The frequency change method, the angle of incidence method, and the group retardation method should give the same equivalent height of the ionized layer.)

R113 Bäumler, M. Feldstärkemessungen auf grosse Entfernungen im Rundfunkwellenbereich. (Field strength measurements at great distances in the broadcasting range). *Elektrische-Nachrichten Technik*, 5, 473–477; November, 1928.

(Report of cooperative field intensity measurements at Königsberg, Hamburg, Karlsruhe, and Dresden. All these stations took readings of waves (190, 405, and 585 meters) arriving in the form of long dashes (30 seconds duration) from a sending station located at the Reichspostamt at Doeberitz. Day and night effects are reported.)

R113 Sreenivasan, K. Über die Wellenausbreitung in einem dispergierenden Medium. (On the wave propagation in a dispersed medium). *Zeits. für Hochfrequenztechnik*, 32, 121–124; October, 1928.

(It is shown that the group velocity of high-frequency waves varies for different frequencies when passing through a dispersed medium such as the Heaviside layer. It seems, therefore, to be evident that a modulated wave on account of its two side bands should produce distortion after passing through this ionized layer.)

R113 Fuchs, J. Der Einfluss der Erdatmosphäre auf die Ausbreitung kurzer Wellen. (On the influence of earth atmosphere on the propagation of short waves). *Zeits. für Hochfrequenztechnik*, 32, 125–129; October, 1928.

(It is shown that the strength of the received signal for short waves after passing over sea water depends on the distribution of the pressure of the atmosphere. From this it follows that the atmosphere produces scattering similar to diffused reflection.)

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R113.5 Goldstein, S. The influence of the earth's magnetic field on electric transmission in the upper atmosphere. *Proc.* Royal Soc. (London), **121A**, 260–285; November, 1928.

(Based on lectures by Prof. J. Larmor. The theory of the effect of the magnetic field of the earth on the propagation of electromagnetic waves in the Heaviside layer is given in much detail.)

R114 Schindelhauer, F. Über elektromagnetische Störungen. (On electromagnetic disturbances). *Elektrische-Nachrichten Technik*, **5**, 442–449; November, 1928.

(Study of the clicks and grinders by means of the direction finder due to Watson-Watt. The author concludes that since the direction of the maximal disturbance is either along or perpendicular to the earth's magnetic axis, most of the atmospherics are due to field changes above the surface of the earth. These field changes cause the electron to be drawn from the sun towards the earth and then produce the eddies of the Heaviside layer. The first causes the clicks and the latter the grinders.)

R114 Watson-Watt, R. A. Present status of knowledge of atmospherics. *Experimental Wireless and Wireless Engineer* (London), **5**, 629–652; November, 1928.

(Reviews work done on this subject by himself and others up to present date.)

R120 Moser, W. Die Übertragung der Energie vom Sender zur Antenne bei kurzen Wellen. (The transfer of energy of short waves from the transmitting set to the antenna). *Elektrische-Nachrichten Technik*, **5**, 422–426; November, 1928.

(Description of the system carrying the high-frequency power to various individual antennas used for beam transmission. The parallel wire and the concentric tube system is used for feeding the power into the antennas and a method is described by means of which the losses of the distributors can be found.)

R125.6 Meissner, A. and Rother, H. Über die Bestimmung des günstigsten Ausstrahlwinkels bei horizontalen Antennen. (On the determination of the favorable radiation angle in horizontal antennas). *Zeits. f. Hochfrequenztechnik*, **32**, 113–115; October, 1928.

(The most favorable radiation angle for 15 and 20 meter wavelengths was determined for horizontal polarization at the center using horizontal multiple antennas in connection with a parabolic reflector. It was found that the most favorable radiation happened when it took place along the tangent of the surface of the earth.)

R125.6 Gresky, G. Die Wirkungsweise von Reflektoren bei kurzen elektrischen Wellen. (The operation of reflectors for short electric waves). *Zeits. fur Hochfrequenztechnik*, **32**, 149–162; November, 1928.

(The beam effect of a vertical antenna for the case of a cylindrical parabolic reflector and a plane reflector (several vertical wires) along a straight wire is experimentally studied. For the parabolic reflector the ratio of focal length to wavelength should be 0.27 and for the plane reflector 0.2. The tuned reflectors give smaller dimensions.)

R125.6 Böhm, O. Die Bündelung der Energie kurzer Wellen. (The concentration of the energy of short waves). *Elektrische-Nachrichten Technik*, **5**, 413–421; November, 1928.

(Explains the beam transmission system employed by the Telefunken Co. A very clear presentation of the underlying principles giving at first the radiation characteristic of the dipole, then that of a group of dipoles along a straight line and in a plane.)

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R125.6 Gothe, A. Über Drahtreflektoren. (On wire reflectors). *Elektrische-Nachrichten Technik,* 5, pp. 427–430; November, 1928.

(Description of the action of reflector antennas. Wire reflectors reduce the strength of the backward beam considerably. Complete screening by means of reflectors is only possible if the radiation coupling between antenna and reflector is variable so that the amplitude and the phase of the reflector current can be properly adjusted.)

R130 Podliasky. Equilibres instables et regimes statiques parasites dans les circuits electriques associes aux triodes. (Unstable equilibrium and regular static parasites in electric circuits associated with tubes). *L'Onde Electrique,* 7, 475–487; November, 1928.

(Conclusion of the paper on pp.287–306 of the July, 1928 issue of this periodical.)

R131 Rajski, C. Les capacities internes de la lampe a plusieurs electrodes. (Interelectrode capacities of multi-electrode tubes). *L'Onde Electrique,* 7, 461–474; November, 1928.

(Expressions are derived for the interelectrode capacities of electron tubes taking the space charge into consideration. If the tube is not burning the usual interelectrode capacities (filament-grid, filament-plate, and grid-plate) are observed but when the filament is emitting electrons it is necessary to consider four capacities, the grid capacity, the plate capacity, the grid-plate capacity, and the plate-grid capacity.)

R134.75 Boella, M. Sur le calcul des amplificateurs a moyenne frequence superheterodyne. (On the calculation of the intermediate frequency amplifier stages of a superheterodyne). *L'Onde Electrique,* 7, 500–508; November, 1928.

(Analytical treatment of the amplifier stages of a superheterodyne used for the amplification of the intermediate frequency.)

R134.75 Ramsay, J. F. A double superheterodyne. *Experimental Wireless and Wireless Engineer* (London), 5, 669–672; December, 1928.

(Description of a two-fold superheterodyne. The first superheterodyne changes the received high frequency to a 600-kc current which is then amplified by two stages of radio-frequency amplification after which another heterodyne produces a 150-kc current. This is passed through three stages of radio-frequency amplification rectified and amplified by a two stage audio-frequency amplifier.)

R144 Waite, G. R., Brickwedde, F. G., Hall, E. L. Electrical resistance and magnetic permeability of iron wire at radio frequencies. *Physical Review,* 32, 967–73; December, 1928.

(Discussion of the results of B. Wwednensky and K. Theodortschik and those of the authors of this paper who could not detect a critical variation in the resistance of iron wire in the vicinity of 3000 kc.)

R144 Jackson, W. The effect of frequency on the value of high resistances of the grid leak type. *Experimental Wireless and Wireless Engineer* (London), 5, 677–679; December, 1928.

(The very high resistance of a grid leak consists in reality of a pure resistance with a small capacity (a few $\mu\mu f$) in parallel. It is shown that above 10^6 cycles/sec. the effective resistance changes and the parasitic capacity current becomes pronounced.)

R200. Radio Measurements and Standardization

R250 Moullin, E. B. An ampere meter for measuring alternating currents of very high frequency. *Proc.* Royal Soc. (London), 121A, 41–71; November, 1928.

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(Gives the theory and construction of a new high-frequency ammeter which is based on the repulsion between two parallel conductors carrying the current to be measured. The frequency effect can be calculated. One conductor is fixed and the other one can move against a small elastic constant. This motion is a measure of the repulsion force and therefore of the current. It is noted by means of a microscope.)

R261 Aiken, C. B. A sensitive vacuum tube voltmeter. *Jnl.* Optical Soc. of American and Review of Sci. Instruments, **17**, pp. 440–450; December, 1928.

(A vacuum-tube voltmeter is described which utilizes the heterodyne principle for obtaining increased sensitivity.)

R300. Radio Apparatus and Equipment

R342.15 Koehler, G. The design of transformers for audio-frequency amplifiers with preassigned characteristics. Proc. I. R. E., **16**, 1742–1770; December, 1928.

(Requirements of ideal transformer are stated and difficulties encountered in attempting to build transformers for interstage coupling units which will meet these requirements are pointed out.)

R343 Küpfmüller, K. Über die Dynamik der selbsttätigen Verstärkungsregler. (On the dynamics of the automatic amplifier stabilizers). *Elektrische-Nachrichten Technik*, **5**, 459–467; November, 1928.

(A system is described by means of which the amplified intensity is automatically kept constant. Based on the principle developed a receiving set has been built which produces the same output intensities during times at which the input voltage (due to fading) varies up and down.)

R344 Eller, K. B. On the variation of generated frequency of a triode oscillator due to changes in filament, current, grid voltage, plate voltage, or external resistance. Proc. I. R. E., **16**, 1706–1728; December, 1928.

(General expressions developed for generated frequency of grid-tuned and plate-tuned generators.)

R344.4 Ritz, M. Essais sur ondes tres courtes. (Tests on very short waves). *L'Onde Electrique*, **7**, 488–499; November, 1928.

(Study of transmission of waves of 2 to 8 meter length. Gives generator diagrams. The experimental results agree with those due to R. Mesny.)

R344.4 Uber eine Methode zur Erzeugung von sehr kurzen elektromagnetischen Wellen. (On a method for the production of short electromagnetic waves). *Zeits. fur Hochfrequenztechnik*, **32**, 172; November, 1928.

(Description of magnetron oscillator for the production of very short waves, $\lambda = 29$ cm.)

R359 Hahnemann, W. Uber die neuere Entwicklung des Maschinensenders fur kleine Wellenlangen. (On the new development of machine transmitters for short wavelengths). *Elektrische-Nachrichten Technik*, **5**, 431–437; November, 1928.

(Description of the latest development of the Lorentz alternators with frequency multiplication. The improvements consist in producing frequencies in the broadcast band; filters for reducing the effect of the side bands; increase of the life of the frequency multipliers and reduction of the Thriller effect which causes a periodic change in the frequency.)

R376.3 Wolff, I. Sound measurements and loudspeaker characteristics. Proc. I. R. E., **16**, 1729–41; December, 1928.

(Description given of methods used to measure loud speaker response. Typica characteristic curves given.)

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R400. Radio Communication Systems

R412 Bailey, A., Dean, S. W., Wintringham, W. T. The receiving system for long wave transatlantic radio telephony. Proc. I. R. E., **16**, 1645–1705; December, 1928.

(Determinations show that frequencies near 60 kc are best suited for transatlantic radio telephone transmission. Various types of antennas described. Mathematical discussions of wave antenna, antenna arrays, and probability of simultaneous occurrence of telegraph interference are given in appendices.)

R500. Applications of Radio

R526.1 Stowell, E. Z. Unidirectional radio beacon for aircraft. Bureau of Standards Journal of Research, December, 1928. Research Paper No. 35. Reprint copies obtainable for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C.

(Description of one of the schemes tried out by the Bureau of Standards for unidirectional radiobeacon work. Directive and non-directive fields are transmitted simultaneously with the proper phase and amplitude relations in order to obtain unidirectional effects.)

R570 Birnbaum, H. W. Die Fernlenkversuche der Reichsmarine in den Jahren 1916–1918. (The guiding experiments of the German Marine in the years 1916 to 1918). *Zeits. fur Hochfrequenztechnik*, **32**, 162–170; November, 1928.

(A description of the system used for guiding airplanes by means of radio.)

R592 Crawley, C. A year's progress in commercial wireless. *Wireless World and Radio Review*, **23**, 801–804; December 12, 1928.

(Discusses automatic S.O.S., position finding, less jamming of broadcast, beam telephony.)

R800. Non-Radio Subjects

534 Ballantine, S. Note on the effect of reflection by the microphone in sound measurements. Proc. I. R. E., **16**, pp. 1639–1644; December, 1928. *Physical Review*, **32**, 988–992; December, 1928.

(Attention is called to the fact that ordinary microphones will not indicate the true pressure of an undisturbed sound wave for the entire audio-frequency band. The correction can, however, be found by employing a standard spherical mounting of which the diagram occupies a small area at the pole. A method of this type can, therefore, be used instead of obtaining the calibration curve with the Raleigh disk.)

534.3 Watanabe, Y. Über die vermittels einer Stimmgabel erregten Rohrenoszillatoren. (On tuning fork vacuum tube oscillators). *Zeits. für Hochfrequenztechnik*, **32**, 116–121; October, 1928.

(The equations for these oscillators are derived and the mechanical as well as electrical oscillations are compared in order to give an expression for the frequency obtained in terms of the true frequency of the tuning fork.)

535.3 Barnard, G. P. The selenium cell: Its properties and applications. *Jnl.* I. E. E. (London), **67**, 97–120; December, 1928.

(Gives the historical review on the work done with the selenium cell and describes the several factors affecting the conductance. Gives applications to photometric and relay problems and shows applications to the optophone, photophone, talking film and television. An extended list of references is given at the end of this paper.)

537.55 Pardue, L. A., Webb, J. S. Ionic oscillations in the glow discharge. *Physical Review*, **32**, 946–49; December, 1928.

(A detailed experimental study of ionic oscillations in the glow discharge which was originally found by Widdington and Appleton.)

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537.65 Tawill, E. P. Nouveau mode de developpement d'electricite par torsion dans les cristaux de quartz. (New method of production of electricity by torsion on quartz crystals). *Comptes Rendus*, **187**, 1042–1044; December **3**, 1928.

(A way was found of producing charges on a quartz cylinder axis along the optical axis when applying a torsion about this axis. Suggests calling it strepho-electricity, because it is different from ordinary piezo-electricity. Shows that for a twist in one direction charges of opposite polarity appear on the surface of the envelope of the cyclinder and the faces perpendicular to the axis. A twist in the opposite direction reverses the polarity. The polarity also depends on the optical rotation.)

537.65 Mandell, W. The determination of the piezo-electric moduli of ammonium Seignette salt. *Proc.* Royal Society (London), **121A**, 130–140; November, 1928.

(Theory and determination of the piezo-electric constants of ammonium Seignette salt.)

621.374.2 Landon, V. D. A bridge circuit for measuring the inductance of coils while passing direct current. Proc. I. R. E., **16**, 1771–1775; December, 1928.

(Bridge circuit described in which inductance of coil is compared to resistances and a capacitance.)

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PROCEEDINGS OF

The Institute of Radio Engineers

Volume 17 April, 1929 Number 4

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The Institute of Radio Engineers

GENERAL INFORMATION

The PROCEEDINGS of the Institute is published monthly and contains papers and discussions thereon submitted for publication or for presentation before meetings of the Institute or its Sections. Payment of the annual dues by a member entitles him to one copy of each number of the PROCEEDINGS issued during the period of his membership.

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Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

Advertising rates to the PROCEEDINGS will be supplied by the Institute's Advertising Department, Room 802, 33 West 39th Street, New York, N. Y.

Changes of address to affect a particular issue must be received at the Institute office not later than the 15th of the month preceding date of issue. That is, a change in mailing address to be effective with the October issue of the PROCEEDINGS must be received by not later than September 15th. Members of the Institute are requested to advise the Secretary of any change in their business connection or title irrespective of change in their mailing address, for the purpose of keeping the Year Book membership catalog up to date.

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Entered as second class matter at the Post Office at Menasha, Wisconsin.

Acceptance for mailing at special rate of postage provided for in the Act of February 28, 1925, embodied in paragraph 4, Section 412, P. L. and R. Authorized October 26, 1927.

Published by

THE INSTITUTE OF RADIO ENGINEERS, INC.

Publication office, 450–454 Ahnaip Street, Menasha, Wis.

BUSINESS, EDITORIAL, AND ADVERTISING OFFICES,

33 West 39th St., New York, N. Y.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume, page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer divisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportion.

Abbreviations—Write a.c. and d.c., kc, μf, $\mu\mu$f, emf, mh, μh, henries, abscissas, antennas. Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

Reprints—With the galley proof a reprint order form is sent to the author. Orders for reprints must be forwarded promptly as type is not held after publication.

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Vice President of the Institute Comes From Germany to Attend Convention

Alexander Meissner
Vice President of the Institute, 1929

The Board of Direction of the Institute is happy to announce that Dr. Alexander Meissner, Vice President of the Institute, is making the trip from Germany to attend the Fourth Annual Convention and to participate in the technical program thereof.

Alexander Meissner was born in Vienna, Austria, on September 14, 1883. He studied at the Vienna College of Engineering and received the degree of Doctor of Technical Science in 1909. In 1922 he was given the honorary degree of Doctor of Engineering by the College of Engineering in Munich. Since 1907 he has been associated continuously with the Telefunken Company of Berlin. He is now Director of the research laboratory of that organization. Dr. Meissner has made a number of fundamental contributions to the radio art through a series of investigations among which are the problems of losses in coils, transmitting antenna design which was especially confined to the study of reducing the resistance losses in the earth, machine transmitters, impulse excitation of quenched spark transmitters, and many vacuum-tube circuits. Dr. Meissner is well known for his active work in the field of vacuum-tube amplification systems, his investigations of quartz crystals in high-frequency work, his study of the physical structure of quartz in piezo and pyro-electricity, and the manufacture of piezo-electric plates from non-crystalline materials.

In 1925 Dr. Meissner was awarded the Heinrich Hertz gold medal in recognition of his many inventions in the radio field.

He became an Associate member of the Institute in 1914 and was transferred to the Fellow grade in 1915. He has been a frequent contributor to the Proceedings and other technical journals. The membership of the Institute elected him Vice President of the Institute in 1929.

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INSTITUTE NEWS AND RADIO NOTES

Washington Convention of the Institute

The Institute is affording its members an unusual program of varied interest at its Fourth Annual Convention which is to be held in Washington, D. C., May 13th to 15th. As well as an excellent technical and inspection program, the Convention committee has arranged a number of special attractions of particular interest to the lady guests to assure all members that their wives are not only welcome but are urged to be in Washington for the entire Convention. The flowers in the many parks of Washington are at their height in May and it is doubtful if they are equalled anywhere in the United States.

Among the features of the Convention will be technical sessions, a popular lecture, technical inspection trips, sightseeing trips for the ladies, a banquet, presentation of the Institute annual awards, a joint meeting of the Institute and the American Section of the International Scientific Radio Union, and a reception by the President of the United States.

Technical Sessions

Thirty-seven papers on radio engineering subjects of timely importance will be presented during the technical sessions. Two half-days will be devoted to symposiums treating the technical problems of radio regulation, and photo radio and television. One full day will be occupied by the joint I. R. E.-U. R. S. I. meeting, at which twenty individual technical contributions will be offered. The *tentative* program of technical papers follows:

MAY 13th, 10:00 A.M. to 12:30 P.M.

Speeches of introduction by President A. Hoyt Taylor, F. P. Guthrie, and C. B. Jolliffe.

Symposium on Technical Problems of Radio Regulation

General Introduction—by the Chairman of the Federal Radio Commission.

Engineering Aspects of the Work of the Federal Radio Commission—by J. H. Dellinger, Federal Radio Commission.

Receiver Developments Affecting Broadcasting Regulations—by L. M. Hull, Radio Frequency Laboratories.

Principles of Broadcast Frequency Allocation—by L. E. Whittemore, American Telephone and Telegraph Company.

Heterodyne Interference—by J. V. L. Hogan, Consulting Engineer.

Ship Radio Inspection and Frequency Monitoring—by Arthur Batcheller, Department of Commerce.

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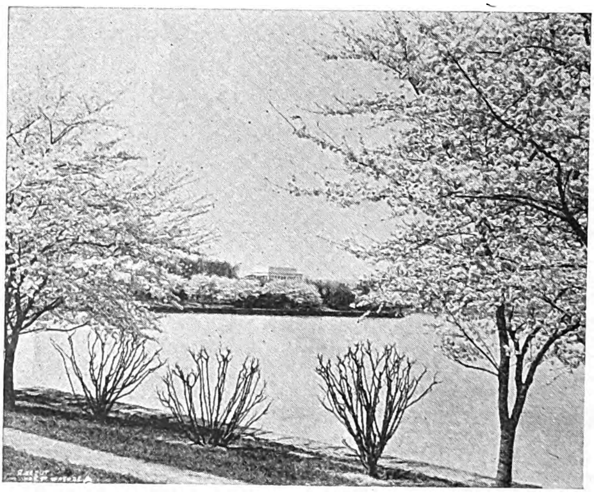
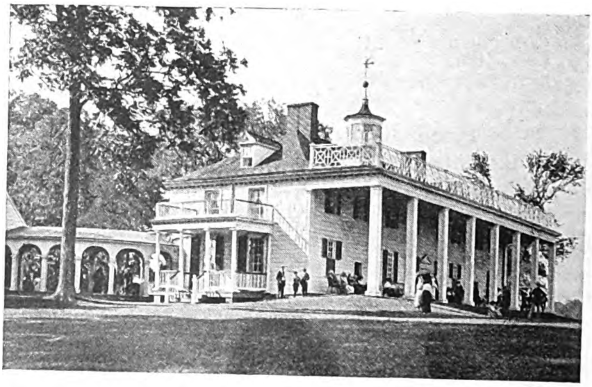
The Capitol

Harris & Ewing



Lincoln Memorial Through the Cherry Blossoms (Trip No. 3)



H. H. Rideout

Mount Vernon (Trip No. 1)

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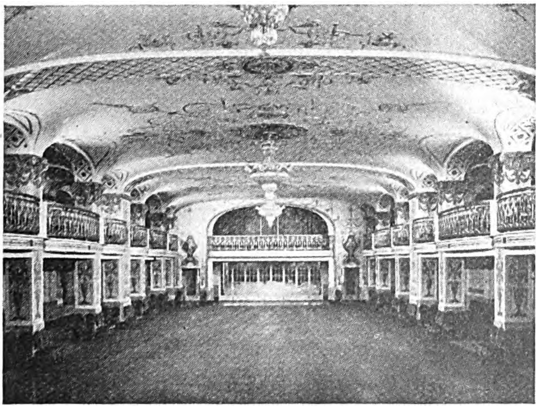
H. H. Rideout

The Japanese Cherry Blossoms—Washington Monument in Background (Trip No. 3)

Promenade of the Mayflower Hotel—Convention Headquarters



Grand Ballroom of the Mayflower Hotel where Annual Banquet Will Be Held

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The Problems Centering About the Measurement of Field Intensity—by S. W. Edwards, Department of Commerce.

Transmitter Developments Affecting Broadcast Regulation—by E. L. Nelson, Bell Telephone Laboratories.

The Radio Engineer's Responsibility in Coping with Man-Made Interference—by Edgar Felix, National Electrical Manufacturers' Association.

Radio Coordination, by M. D. Hooven, National Electric Light Association.

The Development of United States Radio Broadcasting—by R. H. Marriott, Federal Radio Commission.

MAY 14th, 9:00 A.M. to 12:00 M.

Technical Address—by Vice President Alexander Meissner.

Symposium on Photo Radio and Television

General Introduction—by Alfred N. Goldsmith, Radio Corporation of America.

Electrical Transmission of Pictures and Images—by J. W. Horton, General Radio Company.

Commercial Radio Facsimile Communication—by R. H. Ranger, Radio Corporation of America.

High Speed Facsimile Picture Transmission—by V. Zworykin, Westinghouse Electric and Manufacturing Company.

The Drum Scanner in Radiomovies Receivers—by C. Francis Jenkins, Jenkins Laboratories.

The Selection of Standards in Commercial Radio Television—by Julius Weinberger, Theodore A. Smith, and George Rodwin, Radio Corporation of America.

Optical Systems for Television Image Reproduction—by Bernard Cioffari, Radio Corporation of America.

Following the symposium there will be a demonstration of educational sound pictures dealing with the principles of modulation and filtering, presented by John Mills, Bell Telephone Laboratories.

I. R. E.—U. R. S. I. Joint Meeting

MAY 15th, 10:00 A.M. to 5:00 P.M.

Current Developments in Radio Measurement Research—by J. H. Dellinger, Bureau of Standards.

The Inductance of Iron-cored Coils as Determined with Superimposed Alternating and Direct Current, Length of Air Gap, and Initial Magnetic State—by H. M. Turner, Yale University.

Push-Pull Piezo-oscillator Circuits—by J. R. Harrison, Wesleyan University.

The Calculation of the Inductance of Single-layer Coils and Spirals Wound with Wire of Large Cross Section—by F. W. Grover, Union College.

Current Developments in Interference Measurement—by E. F. W. Alexanderson, General Electric Company.

Radio Interference from Power Circuits—by Norman Snyder, General Electric Company.

Current Developments in Radio Wave Propagation Research—by L. W. Austin, Bureau of Standards.

Radio Wave Propagation Phenomena—by Major W. R. Blair, Signal Corps, War Department.

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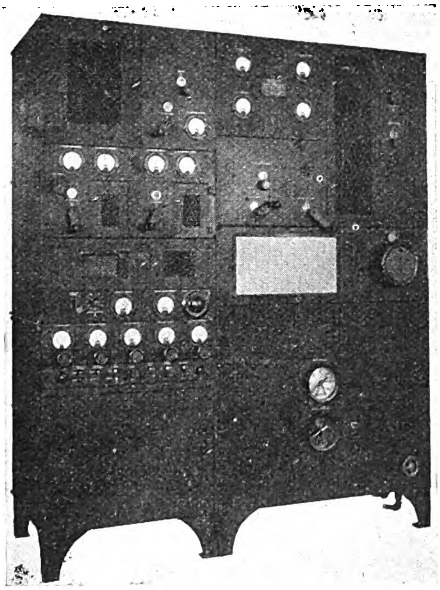
The Effect of Group Velocity on the Retardation of Radio Signals—by G. Breit, Carnegie Institution.

Studies of the Height of the Kennelly-Heaviside Layer by the Echo Method—by M. A. Tuve and L. R. Hafstad, Carnegie Institution.

Further Studies of Echo Signals—by A. Hoyt Taylor and L. C. Young, Naval Research Laboratory.

Ionization in the Atmosphere of Mars—by E. O. Hulburt, Naval Research Laboratory.

Current Developments in Radio Atmospherics Research—by H. T. Friis, Bell Telephone Laboratories.



Front View of Navy High-Power Transmitter
Designed at Bellevue (Trip No. 2)

Current Developments in Radio Research Cooperation—by A. E. Kennelly, Harvard University.

Input and Output Admittance of Triodes—by E. L. Chaffee, Harvard University.

Operating Characteristics of Triode Oscillators—by E. L. Chaffee and R. F. Field, Harvard University.

Reciprocal Theorems in Radio Transmission—by J. R. Carson, American Telephone and Telegraph Company.

Circuit Tuning by Wave Resonance with Application to Signal Reception—by L. Cohen, Signals Corps, War Department.

Application of Wave Resonance Tuning to Signal Transmission—by W. R. Blair and L. Cohen, Signal Corps, War Department.

Detection at High Signal Voltages: I. Plate Rectification—by Stuart Ballantine, Radio Frequency Laboratories.

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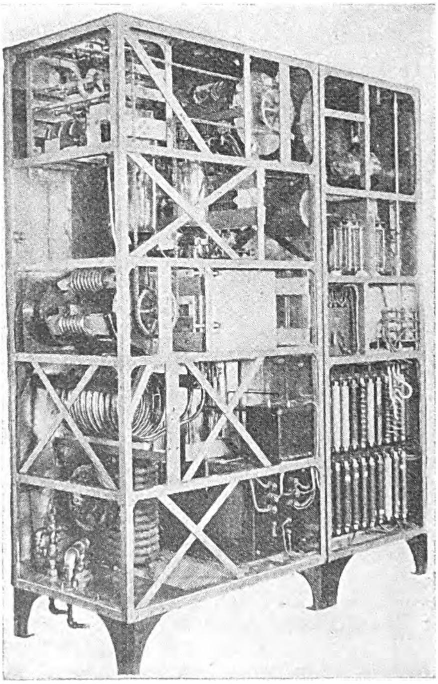
The Mutual Impedance between Adjacent Antennas—by C. R. Englund and A. B. Crawford, Bell Telephone Laboratories.

Reciprocity in Electromagnetic, Mechanical, Acoustical, and Interconnecting Systems—by Stuart Ballantine, Radio Frequency Laboratories.

Popular Lecture

MAY 13TH, 8:00 P.M.

We are particularly fortunate in being able to announce that Professor M. I. Pupin, of Columbia University, past president of both the



Rear View of High-Power Transmitter—Case Removed (Trip No. 2)

Institute of Radio Engineers and the American Institute of Electrical Engineers, will deliver a lecture of general scientific interest.

Dr. Alexander Meissner will preside at this meeting and introduce Professor Pupin.

Inspection Trips

As the places to be inspected on all organized trips are at some distance from the center of Washington (and Convention Headquarters),

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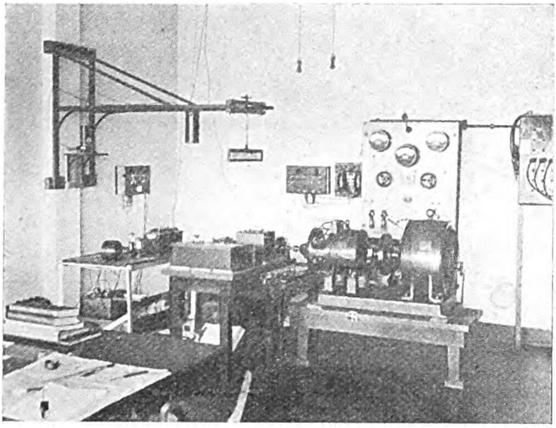
these trips will constitute sightseeing tours as well as technical inspection trips, and consequently will be of much interest to the ladies. Trip No. 1 to Mount Vernon is exclusively for the ladies. The ladies of Washington will be glad to arrange special sightseeing trips for visiting ladies in private cars to points of interest.

Trip No. 2. U. S. Naval Research Laboratory

MAY 13TH, 1:15 P.M. TO 5:30 P.M.

This trip is limited to American citizens; the Naval Research Laboratory is a part of the nation's defense system, and as it is the universal policy among nations to limit visitors to such places to its own citizens, unfortunately this same policy will have to be followed in



One of the Several Forms of Frequency Standardization Equipment at the Naval Research Laboratory (Trip No. 2)

connection with this trip. Members of the Institute and their immediate families planning to take this trip will be required by the Government to register especially for it at the Convention registration desk in the Mayflower Hotel prior to 11:00 A.M. on May 13th.

The Naval Research Laboratory is located seven miles from the center of Washington, at Bellevue, D. C. Buses will start from the Mayflower Hotel and will proceed via the White House and Treasury Building along historic Pennsylvania Avenue past the Botanical Gardens, the Capitol, the Congressional Library, Navy Yard, and St. Elizabeth's Hospital to the Laboratory. Members will return in buses via The Mall, Smithsonian Institute, the National Museum, and the Department of Agriculture, arriving at Convention Headquarters.

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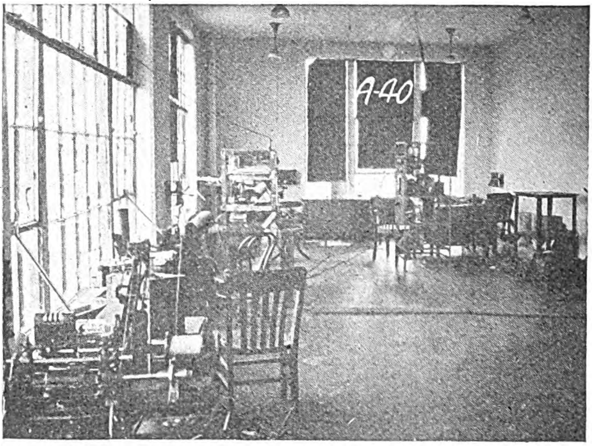
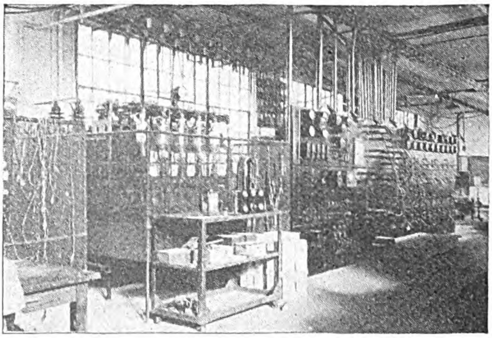
At the Laboratory members will be conducted through the radio and sound divisions. The sections to be visited in the radio division are precision measurement, radio compass, transmitter, aircraft, re-



A Corner of the Naval Research Laboratory's Low-Power Transmitter Design Room (Trip No. 2)

ceiver and amplifier, and lastly the "field house" in which the extensive study of high-frequency transmission phenomena has been carried on for several years.



The High-Power Direct-Current Gallery at the Naval Research Laboratory (Trip No. 2)

The precision measurement section contains the Navy's standards of frequency, permanently installed bridges for inductance, capacity, and resistance measurements, and elaborate equipments for absolute

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calibration of quartz crystals which check back to Observatory time. Of special interest will be the apparatus for absolute calibration by means of which it is hoped an accuracy of one part in ten million will be realized.

The transmitter section will have several problems of general interest in progress. The high-power gallery, in addition to the high-voltage direct-current machinery supply, contains the transmitters with which the Kennelly-Heaviside layer measurements, using the spurt measure-

Bank of Four High-Power High-Frequency Transmitters at Arlington Radio Station (Trip No. 3)

ment method, are being studied. It is expected that a high-power six-meter transmitter will also be in operation.

At the field house the visitors will see a high-power push-pull, high-frequency, crystal-controlled transmitter, as well as a half-kilowatt crystal-controlled transmitter which can be operated up to 35,000 kc. The receiver and oscillograph with which round-the-world signals and echoes are studied will be on display.

The aircraft section will have several sets of historic interest on exhibition. Such equipment as that used on the *Shenandoah*, the

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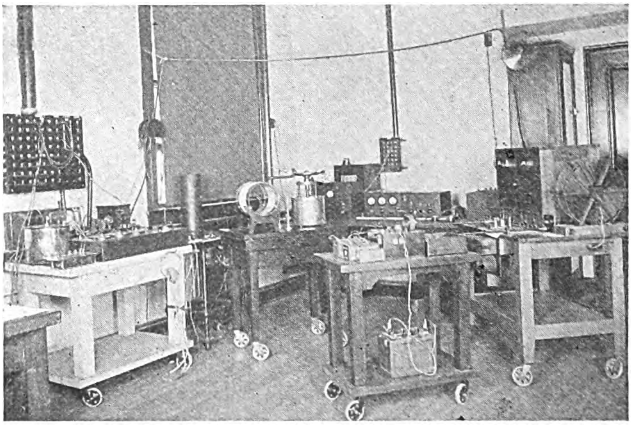
PN7, and Byrd's *America* will be seen. Modern naval aircraft transmitters, receivers, generators, alternators, and dynamotors will be shown. A study of the trend of modern aircraft design, here and abroad, may be made through the comparison of equipment of British and American design on display.

The receiver and amplifier section has a well-equipped laboratory for all manner of receiving investigations. Unique specially screened booths, in which sensitive receiving measurements are made in the midst of high-power high-frequency transmitters, are provided. These booths (three in number) are doubly shielded, each complete screen



Bureau of Standards Laboratory for Testing Piezo-Oscillators for Broadcast Stations (Trip No. 4)

being insulated from the other. They are also arranged so that the center booth can house a driver while in the end one measurement data can be taken without external interference to the indicating instruments, or modulation of the driver frequencies.

Near the main Laboratory there will be found the radio compass building which is designed to approximate the standard shore station layout. In this building the radio compass section carries on its investigations and development work.

At the docks on the Potomac river will be seen the field stations of the sound division—consisting of two well equipped barges which are operated in Chesapeake Bay except during the winter months. The sound division will have on display a direct reading sonic depth finder. There will also be a demonstration of a few applications of piezo-electric activity of Rochelle Salt crystals.

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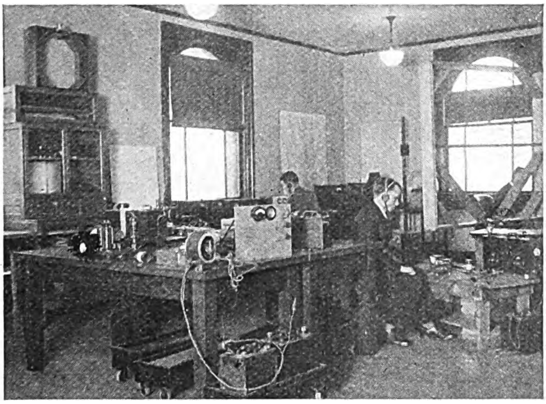
Trip No. 3. Sightseeing with Stops at Arlington Cemetery and Arlington Radio Station

May 14th, 10:00 a.m. to 12:30 p.m.

This trip will parallel a portion of the symposium on photo radio. Leaving the Mayflower in buses, those participating in this trip will be taken on a two and a half-hour trip to many famous points of Washington. Lecturers will be provided on each bus. The buses will pass the War and Navy Buildings, Corcoran Art Gallery, Red Cross Building, Pan American Union, National Academy of Sciences, Lincoln Memorial, around Haynes Point from which an excellent view of the Naval Air Station, Bolling Field, and the War College may be had, past the



Laboratory for Study of Radio Wave Phenomena and Measurement of Intensity of Radio Waves at the Bureau of Standards (Trip No. 4)

public golf links, the Bureau of Printing and Engraving, the Tidal Basin surrounding which are the famous Japanese Cherry trees, to Arlington Cemetery and the Tomb of the Unknown Soldier. From Arlington Cemetery the buses will go to Arlington Radio Station, well known to all, where a stop will be made to inspect the latest low- and high-frequency naval transmitters in operation. The return trip will be through Fort Meyer across Francis Scott Key bridge, through historic Georgetown to the Mayflower Hotel.

Trip No. 4. U. S. Bureau of Standards

May 14th, 1:15 p.m. to 5:30 p.m.

En route the Bureau of Standards, buses with lecturers will leave Convention Headquarters, proceeding up 16th Street, passing many embassies and fine churches, through the finest residential section of Washington to one of the north entrances of Rock Creek Park. A

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four-mile ride through this famous picturesque park will bring the delegates out at Old Pierce Mill. From thence the route will lead direct to the Bureau of Standards. The return trip will be made along Connecticut Avenue, across the Million Dollar bridge, past the former home of President Hoover and the "S" Street home of Woodrow Wilson to the Mayflower hotel.

The Bureau of Standards affords a large number of interesting sights. The attention of the members will be concentrated on the radio and electrical buildings, but opportunity will be given for anyone

Radio Transmission Research (Low-Frequency) Laboratory of Dr. L. W. Austin at the Bureau of Standards (Trip No. 4)

interested to visit other laboratories. The work of the Bureau of Standards includes nearly all phases of physics, chemistry, and engineering and informal visits to any section will be welcomed. Information may be obtained from the laboratory members.

In the radio building there are many interesting developments to be observed. Members of the staff will be available for answering questions and discussing the various projects. Probably of most interest will be the following: the study of radio wave phenomena and measurement of radio field intensity showing apparatus for recording fading and illustrative records, the laboratory car and other measuring equipment, demonstration of testing of piezo-oscillators for broadcast stations and measurements on standard piezo-oscillators, demonstration of the visual radio beacon system and display of apparatus developed in connection with work on radio aids to air navigation, the laboratory of Dr. L. W. Austin for radio transmission research (low frequency) where the classic transmission measurement work of Dr. Austin, which the Proceedings has published from time to time, has

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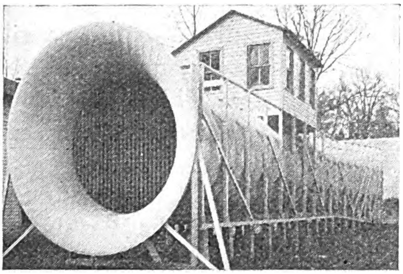
been carried on for a number of years, the Signal Corps Laboratory where the work on resonance wave tunings will be demonstrated.

In the Electrical Building will be seen the life test equipment for incadescent lamps where several thousand lamps are tested annually, heavy current testing and tests of electrical measuring instruments, standard cell laboratory in which is maintained the United States standard of voltage and life testing of dry batteries, including radio batteries in which all switching is done automatically according to any predetermined time schedule.



Ten-Foot Wind Tunnel for the Study of Effect of Wind on Structures Such as Airplanes, Dirigibles, and Buildings, at the Bureau of Standards (Trip No. 4)

In the main laboratory of the Wind Tunnel Building there is one of the three wind tunnels used at the Bureau for investigating models of airplanes, dirigibles, bombs, buildings, and chimneys. This tunnel is fifty inches across at its smallest section, and in it a wind speed of 75 miles an hour can be maintained.

The work of the Bureau on radio aids to air navigation is carried on at an experimental station at College Park, Maryland. Members interested in this work are welcome to visit the station at any time between 9:00 A.M. and 4:30 P.M. Special arrangements for a visit by a group of members may be made at Convention Headquarters at registration time.

Reception By President Hoover

A reception by the President of the United States at the White House to all persons attending the Convention is being arranged. The date and hour will be announced later.

The Banquet

MAY 14TH, 7:30 P.M.

The annual banquet of the Institute, which will be informal, will be held in the grand ballroom of the Mayflower Hotel. Included in

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the program are speeches by A. Hoyt Taylor, toastmaster, and by several other members of the Institute, presentation of the Institute Medal of Honor and the Morris Liebmann Memorial Prize, and several forms of light entertainment. Dancing will be provided for those who desire.

Entertainment for the Ladies

From the foregoing paragraphs it may be judged that the lady guests are to be provided with a varied program of interest. In addition to these the Ladies' Committee of the Convention has arranged a number of added attractions.

On Monday morning the Ladies' Committee will be present at the registration desks to arrange informal sightseeing trips exclusively for the ladies. These trips will be made in private cars and will include any of the sights of Washington which are of interest to the individual persons.

At noon on Monday the ladies of Washington will be hostesses to the lady guests at luncheon at one of the Washington Women's Clubs. Following the luncheon the ladies will go on Trip No. 1 to Mount Vernon. This will be a four-hour tour by motor-bus and will include Georgetown, Alexandria, and Mount Vernon.

The Ladies' Committee will be on hand at other times to direct women visitors to any of the numerous interesting sights which they may care to take in.

Committee on Sections Meeting

May 15th, 6:30 p.m.

A meeting of the Committee on Sections, which it is believed will be attended by delegates from each Section of the Institute, will be held at an informal dinner at the National Press Club, 14th and F Streets, N. W. Those members of the Sections who attended this session of the 1928 Convention will recall the helpful and instructive meeting which took place. These meetings afford the only opportunity for the Officers of the Institute and the Committee on Sections with its various representatives to meet in a body to study Section plans and operation. The 1929 Committee on Sections is arranging a very interesting program for this meeting.

General Information

Convention Headquarters will be at the Mayflower Hotel which is centrally located (Connecticut Ave. and L St. N. W.), and very near the U. S. Chamber of Commerce building in which the technical sessions will be held.

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The Mayflower Hotel has set aside the following rooms for Convention delegates: Sixty-five single rooms at $5.00 a day, twenty-five single rooms at $6.00 a day, twenty-five rooms at $7.00 a day, fifty twin-bedded rooms at $9.00 a day for two and $7.00 for one, fifty twin-bedded rooms at $10.00 for two and $8.00 for one. If more than one person occupies a single room all of which are provided with double beds the price will be $2.00 a day additional. To insure reservations desired the members are cautioned to fill in and return promptly the hotel reservation postcard which will be sent with the Convention programs during the latter part of April.

Members coming to Washington by auto will find a number of garages near the Mayflower. Among these are: Mayflower Garage, 1830 L Street, $1.50 a day; Standard Garage, 1815 L Street, $1.00 a day; L Street garage, 1705 L Street, $1.00 a day; and the Palace Garage, 1216 20th Street, $1.00 a day.

Also with the Convention programs there will be mailed a detailed map of the city of Washington showing the auto approaches to the city and Convention Headquarters.

Tickets for all inspection trips must be obtained at the time of registration either on Sunday, May 12th or Monday, May 13th, prior to the opening session. These tickets will be available free of charge to members of the Institute and their immediate families. Due to the limited facilities it will be impossible to take non-members, other than their families, on any inspection trip or sight-seeing tour.

The banquet tickets will be $5.00 a plate, and may be obtained either at the time of registration or in advance thereof from Mr. F. P. Guthrie, Radio Corporation of America, 1112 Connecticut Avenue, Washington, D. C.; or the Secretary, The Institute of Radio Engineers, 33 West 39th Street, New York City. Remittances must accompany orders for tickets. Particular table assignments will be made only when one or more complete tables (seating ten persons) are reserved.

March Meeting of the Board of Direction

The March meeting of the Board of Direction of the Institute was held on the 6th of the month in the office of the Institute. The following were present: R. H. Marriott, Acting Chairman; Melville Eastham, Treasurer; John M. Clayton, Secretary; Alfred N. Goldsmith, Junior Past President; Arthur Batcheller, W. G. Cady, R. A. Heising, and R. H. Manson.

The Board transferred and elected the following to the higher grades of membership in the Institute: transferred to the Fellow grade:

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Haraden Pratt; elected to the Fellow grade: Guiseppe Pession; transferred to the Member grade: W. D. Loughlin; elected to the Member grade: Roland B. Bourne, G. A. Struthers, and Ralph Wyckoff.

One hundred and twenty-one Associate members and eight Junior members were elected.

Standard Frequency Transmissions by the Bureau of Standards

The schedule of standard frequency transmissions by the Bureau of Standards for the months of April and July (inclusive) appear below. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington which went into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to 1,000 miles from the transmitting station.

The transmissions are by continuous-wave radiotelegraphy. The modulation which was previously on these signals has been eliminated. A complete frequency transmission includes a "general call" and "standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues for about 2 minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about 4 minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

Radio Transmissions of Standard Frequency; Schedule of Frequencies in Kilocycles

Eastern Standard Time	March 20	April 22	May 20	June 20	July 22
10:00–10:08 P.M.	1500	4000	125	550	1500
10:12–10:20	1700	4500	150	600	1700
10:24–10:32	2250	5000	200	700	2000
10:36–10:44	2750	5500	250	800	2300
10:48–10:56	2850	6000	300	1000	2700
11:00–11:08	3200	6500	375	1200	3100
11:12–11:20	3500	7000	450	1400	3500
11:24–11:32	4000	7300	550	1500	4000

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Institute Meetings

Atlanta Section

The Atlanta section held a meeting February 8th in the Federal Building, Atlanta, Georgia, presided over by W. Van Nostrand, chairman. There was an attendance of twenty-seven members.

J. K. Clapp, of the General Radio Company, presented a paper entitled, "A Convenient Method for Referring Secondary Frequency Standards to a Standard Time Interval." This paper was published in the February, 1929, issue of the Proceedings.

Buffalo-Niagara Section

The February meeting of the Buffalo-Niagara section was held at the University of Buffalo, Buffalo, N. Y., on February 14th. L. C. F. Horle, chairman, presided. Fifty members attended.

The following papers were presented: "Gain Measurements at Amateur Band Frequencies," by F. A. Lidbury; "Resistance Measurements at Broadcast Frequencies" by Alton Velia; and "Design and Construction of Beat Frequency Oscillator," by G. Edgar Stone. A general discussion followed the presentation of these papers.

Cleveland Section

The Cleveland section held a meeting in the Case School of Applied Science, Cleveland, Ohio, on February 28th. Bruce W. David, chairman of the section, presided. Fifty-five members attended.

O. T. McIlvaine presented a paper, "Vacuum Tubes—New Developments and Problems," describing a new rectifier tube which converts 110-volt alternating current to 220-volt direct current without a transformer. Two separate anodes and heater type cathodes are enclosed in the tube which fits a standard 227-type socket. The heater elements are supplied from a 6-volt source of alternating current or direct current. A general discussion followed the presentation of this paper.

Los Angeles Section

On February 18th the Los Angeles section held a meeting in the Elite Cafe, Los Angeles, California. T. F. McDonough, chairman, presided. Sixty-four members attended the meeting.

R. B. Parrish and G. D. Ellis presented short informal talks on "Sound Recording As Applied To Motion Pictures" and "Photophone System of Sound Recording." A general discussion followed.

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New York Meeting

On March 6th a New York meeting of the Institute was held in the Engineering Societies Building, 33 West 39th Street. Alfred N. Goldsmith, Past President, presided in the absence of President Taylor.

G. L. Beers, of the Westinghouse Electric and Manufacturing Company, presented a paper by W. L. Carlson and himself entitled "Recent Developments in Superheterodyne Receivers."

Following the presentation of this paper, which was published in the March, 1929, issue of the Proceedings, it was discussed by Messrs. Beers, Goldsmith, MacDonald, Jones, Murray, and others. Four hundred and twenty-five members of the Institute and their guests attended this meeting.

Philadelphia Section

The Philadelphia section held a meeting on February 19th at the Franklin Institute, Philadelphia, Pa. J. C. Van Horn, chairman, presided. Twenty members of the section attended the meeting. J. K. Clapp, of the General Radio Company, presented a paper, "A Convenient Method for Referring Secondary Frequency Standards To A Standard Time Interval." This paper was published in the February, 1929, issue of the Proceedings.

Messrs. Miller, Synder, Van Horn, and others participated in the discussion which followed.

Pittsburgh Section

J. K. Clapp, of the General Radio Company, was the speaker at the Pittsburgh section meeting, which was held February 21st in the Fort Pitt Hotel, Pittsburgh, Pa. W. K. Thomas presided and twenty-two members of the section attended.

The paper, "A Convenient Method for Referring Secondary Frequency Standards to a Standard Time Interval," which was presented by J. K. Clapp, was published in the February issue of the Proceedings.

Messrs. McKinley, Hitchcock, Harmon, Mag, Thomas, and Allen participated in the discussion which followed.

San Francisco Section

The San Francisco section held a meeting February 13th in the Bellevue Hotel, San Francisco, Calif. Donald K. Lippincott, chairman, presided and fifty-one members of the Institute attended.

Haraden Pratt, Chief Engineer of the Mackay Radio and Telegraph Co., presented a paper, "Radio Communication in the Field of Aero-

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nautics." The paper told of the development of beacon reception on airplanes, starting with the trailing wire antenna. The difficulties with this type of antenna were told, and the inaccuracies in direction pointed out. A description of the transmitters used to give beacon signals was given, and the final development of a two-reed indication receiver was explained. Field intensity curves for the beacon transmitters were shown, and the variation of this curve by variation of phase angle between two loops or transmitters was explained. The vertical pole antennas were described as being very true in directional qualities, and in utilizing the field of the propagated wave night effects were eliminated. The shielding of ignition, including spark plugs, was described. A general discussion followed the presentation of this paper.

Chairman Donald K. Lippincott announced the appointment of the Publicity, Meetings and Papers, and Membership Committees. Arthur H. Halloran was appointed chairman of the Publicity Committee, and Preston Allen and John L. Tustin, members; Benjamin E. Wolf, chairman of Meetings and Papers Committee, with George T. Royden and Philo T. Farnsworth, members; and Bernard H. Linden, chairman of the Membership Committee, with William P. Bear and Clifford J. Dow, members.

Toronto Section

On February 20th the Toronto section held a meeting in the Electrical Bldg., University of Toronto, Toronto, Canada. A. M. Patience, chairman of the section, presided. Fifty-four members attended.

J. H. Thompson, Chief Engineer of the Canadian Marconi Co., Montreal, presented a paper, "Marconi Beam System." The system of transmission and reception were explained, and methods of feeding the aerial and of concentrating and directing the beams were explained in detail. The illustrated slides showed the Beam Stations at Drummondville and Yamachiche, Province of Quebec. Messrs. Mott, Roseburgh, Patience, and others participated in the discussion which followed.

Washington Section

The Washington section held a meeting February 14th in the Continental Hotel, Washington, D. C. F. P. Guthrie, chairman, presided. Sixty-five members and guests attended the meeting.

J. K. Clapp, of the General Radio Company, presented a paper, " A Convenient Method of Referring Secondary Frequency Standards to a Standard Time Interval." This paper was published in the

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February, 1929, issue of the PROCEEDINGS. August Hund and others entered into the discussion which followed.

Committee Work

COMMITTEE ON ADMISSIONS

At the March 6th meeting of the Committee on Admissions the following members were present: R. A. Heising, chairman; H. F. Dart, A. F. Van Dyck, E. R. Shute, J. S. Smith, and B. W. David, chairman of the Cleveland Section. The Committee considered thirty-nine applications for transfer or election to the higher grades of membership in the Institute.

COMMITTEE ON MEMBERSHIP

A meeting of the Committee on Membership was held in the Institute office on March 6th with attendance as follows: H. B. Coxhead, acting chairman; J. C. Stroebel, Jr., A. F. Murray, and F. R. Brick.

The Committee considered in detail a memorandum circulated by Mr. Coggeshall, chairman, outlining a very progressive program of work for 1929.

The Committee particularly solicits correspondence from out of town members to secure the benefit of the points of view of members throughout the entire world. It is felt that the growth of the Institute's membership can be materially assisted if members will endeavor to bring the Institute and its aims and activities before prospective members whenever an occasion presents itself.

COMMITTEE ON SECTIONS

The first meeting of the 1929 Committee on Sections was held in the Institute Office at 5:00 P.M. on Wednesday, March 6th. The following were present: E. R. Shute, chairman, Austin Bailey, L. A. Briggs, A. F. Murray, Q. A. Brackett (of the Connecticut Valley Section), B. W. David (of the Cleveland Section), and C. J. Porter, Assistant Secretary.

The Committee discussed the Section meeting program for the 1929 Convention. The Committee initiated the revision of the "Section Manual" which was prepared several years ago for the benefit of persons interested in the organization of Sections.

In detail the Committee considered correspondence with persons in five territories in which it was thought that Sections might be

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organized. It is anticipated that the majority of these prospective Sections will be in operation before the end of the year.

Committee on Meetings and Papers

A meeting of the Committee on Meetings and Papers was held at 9:30 A.M. March 6th in the Institute Office, 33 West 29th Street. Those present were K. S. Van Dyke, chairman; W. G. Cady, Wilson Aull, Ralph Batcher, Edgar Felix, Virgil Graham, J. W. Horton, W. H. Murphy, Greenleaf W. Pickard, R. H. Ranger, Julius Weinberger, W. Wilson, B. E. Shackelford, W. C. White, and H. F. Olson.

The Committee discussed the relation between the Board of Editors and the Committee on Meetings and Papers. A subcommittee to prepare a code of the Committee on Meetings and Papers was appointed. This Committee personnel is W. C. White, chairman; B. E. Shackelford, W. G. Cady, and K. S. Van Dyke.

To arrange for the programs of New York meetings of the Institute a subcommittee of the Committee on Meetings and Papers was appointed as follows: W. Wilson, chairman; F. H. Kroger and John M. Clayton.

The Committee read and approved the tentative program of the technical portion of the 1929 Convention.

Correction

Through typographical error in connection with the publication of the paper "Reception Experiments in Mt. Royal Tunnel" in the February, 1929, issue of the PROCEEDINGS, the positions held by two of the authors were incorrectly listed. A. R. McEwan should be listed as Director of Radio Department of the Canadian National Railways; G. W. Olive as Radio Engineer, Department of Radio, Canadian National Railways; and J. H. Thompson as Chief Engineer of the Canadian Marconi Company.

Personal Mention

William K. Aughenbaugh, formerly operating engineer at Station WFBG at Altoona, Pa., has recently become control room operator of Station WLW in Cincinnati, Ohio.

L. F. Jones, 3d, has been transferred from the Schenectady Radio Engineering Department of the General Electric Company to the same department of the General Electric Company at Oakland, Cal. Mr. Jones is engaged in short-wave development work.

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Louis E. Kearney has recently resigned from the Photo Radio Laboratory of the Radio Corporation of America at New York City to become Radio Inspector for the Port of Philadelphia, replacing John E. Leitch, who is now Chief Engineer for Station WCAU of the Universal Broadcasting Company.

Paul J. Schwarzhaupt, formerly with the Radio Department of Spear and Company of New York City, is now in the Radio Engineering Department of the General Electric Company at Schenectady, N. Y.

Elmer L. Brown has resigned from the California Victor Distributing Company to assume a position with the Engineering Division of the Technical and Test Department of the Radio Corporation of America in New York City.

Leonard F. Fuller, member of the Board of Direction of the Institute in 1928 and formerly Chairman of the San Francisco Section, has resigned from the General Electric Company at San Francisco, where he was engaged in carrier current work and application of high-voltage high-frequency equipment in the light and power industry, to become executive vice-president of the Federal Telegraph Company in charge of that organization's affairs on the west coast. Dr. Fuller's headquarters are at Palo Alto.

Dan T. Fernandez has left the Stewart Warner Speedometer Company of Chicago to become audio engineer for the Day Fan Electric Company of Dayton, Ohio.

L. T. Newell, of the U. S. Lighthouse Service, has been transferred from Washington, D. C. to the Airways Communication Station at Bellefonte, Pa., as operator in charge.

O. H. Caldwell, formerly a member of the Federal Radio Commission, has returned to New York City to resume his position as Editor of *Radio Retailing* and *Electrical Merchandising*.

Robert Hertzberg, for the past two and one-half years managing editor of *Radio News* magazine, has resigned that position to join the staff of the Pilot Electrical Manufacturing Company of Brooklyn.

I. G. Maloff, formerly associated with John Minton at White Plains, N. Y. in consulting work, is now vice-president and in charge of engineering for Valley Appliances, Inc., of Rochester, N. Y.

Carrington H. Stone is now associated with Jenkins and Adair of Chicago, Ill., working on talking motion picture and public address system problems. Mr. Stone was formerly assistant chief engineer of the Radio Division, Stewart-Warner Speedometer Corporation, Chicago, Ill.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED MARCH 6, 1929

Transferred to the Fellow Grade

New York	New York City, 67 Broad Street	Pratt, Haraden

Elected to the Fellow Grade

Italy	Rome, Via Tevere No. 20	Pession, Guiseppe

Transferred to the Member Grade

New Jersey	Boonton, Radio Frequency Laboratories	Loughlin, W. D.

Elected to the Member Grade

Connecticut	Hartford, 221 Holcomb St.	Bourne, Roland B.
Oklahoma	Ponca City, 604 Marland Drive	Wyckoff, Ralph D.
England	Cornwall, Bodmin, Radio Beam Station	Struthers, G. A.

Elected to the Associate Grade

Alabama	Mobile, c/o Adams Glass and Co., Inc.	Blakeney, George H., Jr.
Arkansas	Little Rock, 214 West 4th Street	Bilheimer, Joe Allen
California	Burbank, 320 No. Tujunga Ave.	Hernfeld, Frank P.
	Glendale, 202 East Broadway	Mitchell, G. A.
	Monterey, 460 Alvarado Street	Newman, Clarence D.
	Oakland, 1609 Grand Avenue	Wild, Sidney J.
	Pasadena, 230 South Hudson Avenue	Taylor, Kenneth
	San Francisco, 328 Custom House	Kellogg, Frank L.
	San Francisco, 9 Duncan Street	Pinkney, J. L.
	San Pedro, USS California, Box 17, c/o Postmaster	Hart, Clyde Armand
	Stockton, 543 So. California Street	Green, Alfred H.
Connecticut	New Haven, 104 Lake Place	Gounarides, A. D.
	New Haven, 1523 Chapel Street	Rhoades, A. L.
	Waterbury, Y. M. C. A.	Knowlton, John J.
Dist. of Columbia	Washington, Radio Laboratory, Bureau of Standards	Doherty, William H.
	Washington U. S. Naval Research Laboratory	Wallace, James D.
Florida	Fort Myers, 1138 Franklin Street	Johnson, Carl C.
	Jupiter, Box 158	Miller, Joseph Anthony
	Winter Park, Hamilton Hotel	Emerson, John
Georgia	Atlanta, 912 Candler Bldg.	Hobart, James C.
	Atlanta, Box 1593	McKinney, Joe H.
	Fort Benning, P. O. Box 615	Sosbee, Frank E.
Illinois	Brookfield, 126 So. Vernon Avenue	Courter, H. L.
	Chicago, 4849 No. Lawndale Avenue	Eppstein, Ralph M.
	Chicago, 4514 B. N. Ashland Avenue	Frederick, William J.
	Chicago, 1450 Argyle Street	Hitzenhammer, Anthony de V.
	Chicago, 4747 Kenmore	Riggs, John T.
	Wilmette, 1025 Sheridan Road	Freis, J. H.
Louisiana	St. Francis, P. O. Box 232	Harris, David H.
Maine	Houlton, P. O. Box 69	Willes, Emerson T.
	Portland, 56 Lafayette Street	Giro, Daniel J.
Maryland	Baltimore, 1611 Lexington Bldg.	Mathers, Earl S.
Massachusetts	Boston, 160 W. Brookling Street	Nelson, Robert Neil
	Cambridge, 57 Gorham Street	Shen, S. Z.
	Northampton, 126 Franklin Street	Mettey, Bradford Joseph
	Quincy, 100 Bay View Avenue	Renton, Ralph James
	South Dartmouth, M. I. T., Round Hills	Houghton, Henry G., Jr.
Michigan	Ann Arbor, 304 S. Fifth Avenue	Allen, Roy B.
	Ann Arbor, 423 Fuller Street	Embury, Milburn S.
	Ann Arbor, 402 N. Main St.	MacCaffray, R. S., Jr.
	Ann Arbor, 403 S. Fifth Avenue	Martin, J. Faber
	Ann Arbor, 109 Packard Street	Pickett, Willis N.
	Ann Arbor, 1666 Broadway	Richmond, Clifford A.
	Ann Arbor, 1108 Michigan Avenue	Wilber, Harold
	Birmingham, 33 Linden Road	Soeters, Raymond A.
	Detroit, 3403 Cicotte	Tomchuck, John
	Fordson, 5615 Horger Avenue	Knight, Donald M.
	Oxford	Capron, E. S.
Missouri	St. Louis, 3727 Juniata Street	Ruddy, Ralph P.
	University City, 8224 Fairham Avenue	Tevis, Graham L.
New Jersey	East Orange, 5 Eppirt Street	Somers, Richard M.
	Jersey City, 241 Van Vorst Street	Benin, Zolmon
	Keanburg, 389 Palmer Avenue	Herrmann, Edwin
	Orange, 459 Hepwood Avenue	Harris, Gwin C.
New York	Brooklyn, 400 Lincoln Place	Rojas, Fernando R.
	Jamaica, 12 Allen Street	Yenoli, Dominick J.
	New York City, City Island, Harlem Yacht Club	Appel, Henry W.
	New York City, 1160 Bryant Ave., Bronx	Atkin, Robert
	New York City, 2510 Davidson Ave., Bronx	Bond, Elmer Frederick

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New York (cont'd)	New York City, 411 Fifth Avenue	Frank, James, Jr.
	New York City, R. C. A., 70 Van Cortlandt Park So.	Georghegan, Eamonn
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Part II

TECHNICAL PAPERS

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

FREQUENCY MULTIPLICATION BY SHOCK EXCITATION*

By
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([1]Department of Electrical Engineering, Massachusetts Institute of Technology, Cambridge, Mass.
[2]Department of Electrical Engineering, University of Missouri, Columbia, Mo.)

Summary—*The fundamental principles involved in the theory of frequency multiplication by means of iron-core coupled circuits are briefly reviewed from the standpoint of Fourier analysis as well as that of recurring transients. Oscillograms and figures illustrating the effect of transformer inductance and shock duration upon wave form are given. Oscillations of the first and second kind are discussed in their relation to the analogous arc circuit problem. Efficiency and power output are considered in their dependence upon primary and secondary current amplitude, and the conditions for smoothest wave form and maximum efficiency are pointed out.*

THE method of frequency multiplication discussed in this paper makes use of the fact that if a non-harmonic voltage is impressed upon a simple oscillatory circuit, the response will reach a maximum whenever the natural frequency of the circuit coincides with one of the harmonics contained in the impressed voltage. If the latter is highly distorted, i.e., contains large harmonic components, then the response of the oscillatory circuit will of course also be appreciable. The circuit arrangement which produces this distorted voltage and impresses it upon an oscillatory mesh is given in Fig. 1. The transformer inductance L_t is small compared with L_1. Hence if $L_1$and C_1 are tuned to the machine frequency, the primary circuit will be practically in resonance, and the current will be nearly sinusoidal regardless of the saturation effects in the transformer. The latter is so designed that saturation sets in at a current value considerably below the primary maximum, so that the induced voltage is highly distorted. This is the voltage which is impressed upon the secondary tuned circuit, thus giving rise to frequency multiplication.

This arrangement has been investigated both theoretically[1] and experimentally in the past, but the present paper proposes to throw light upon some of the features involved which have not yet been fully appreciated. Some of these are: the effect of the duration of shock upon response and wave form; the production of frequencies which are even multiples of the fundamental; oscillations of the first and second kind and their analogy to the case of frequency multiplication by means of arc circuits; the effect of shock duration upon efficiency; etc.

* Dewey decimal classification: R146. Original manuscript received by the Institute, December 19, 1928.

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A rigorous mathematical theory based upon assumptions elaborated below has been given[1] for this type of circuit, and consequently will not be reproduced in its entirety here. We shall rather limit ourselves to a brief review of the method of attack and some of its consequences for the purpose of gaining for the reader a better orientation in the present picture so that he may follow without difficulty the arguments which are presented to clarify the particular phenomena offered in this paper.

We have called this method of frequency multiplication (following the suggestion of other writers on this subject) the method of *shock excitation.* It might also be called the method of *filtering harmonics,* as will easily be appreciated by the reader from the remarks made in the opening paragraph. The difference in these two ways of designating the same phenomena lies in a fundamental difference in the mathematical method of attack, of course, and has no counterpart in the actual physical mechanism involved. Since each way of looking upon the situation possesses some distinct advantages in the nature of illuminating certain features of the general problem, both will be presented below, at least in fundamental outline.

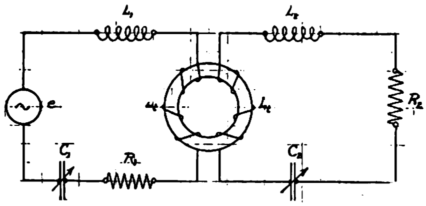

Fig. 1—General Circuit Arrangement.

The basic simplifying assumptions, by which the mathematical attack is made possible, are the same for both methods. First we idealize the magnetization curve as shown in Fig. 2, i.e., the slope is assumed constant between the saturation points, and zero above these. Although this rather strong idealization may seem to depart too radically from actual fact, it nevertheless retains (and even accentuates) those features which are the very essence of our problem, thereby amplifying the points of chief interest. Our second basic assumption is that the primary current varies linearly with time throughout the interval in which the transformer is below saturation. Since the primary current is sinusoidal, and its maximum large compared with i_{sat}, this is nothing more than replacing sin x by x for small values of the latter.

[1] "Zur Theorie der Frequenzvervielfachung durch Eisenkernkoppelung," *Archiv. für Elektrotechnik,* Band XVII, 1926, Heft 1, S. 17.

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Our first assumption obviously causes the induced voltage in the transformer to be zero unless the flux is below the saturation value, and the second assumption gives this induced voltage a constant value during the period below saturation. This induced voltage therefore, which is the internally generated voltage for the secondary circuit, takes the

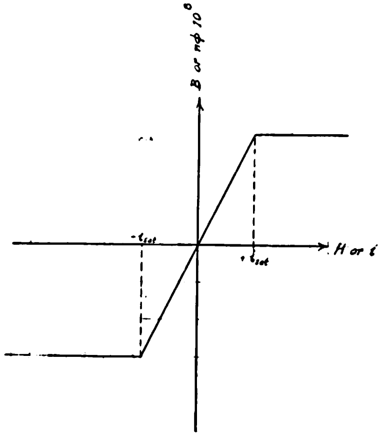

Fig. 2—Idealized *B-H* Relation.

form shown in Fig. 3. It is a periodic succession of rectangular impulses with the same period as that of the generator in the primary circuit.

This internally generated voltage is impressed upon a very peculiar circuit, however, and it is just this peculiarity which accounts for numerous anomalies to be pointed out later; namely, the inductance of the secondary circuit is not a constant, but equal to L_2+L_t so long as

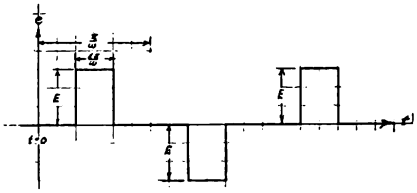

Fig. 3—Resulting Induced Transformer Voltage Plotted with Symmetrical Time Origin for Simplification of Fourier Expansion.

the transformer is below saturation, and L_2 above saturation, because here L_t is zero due to the zero slope of the magnetization curve. These discontinuities in the secondary circuit inductance coincide in time with those of the voltage of Fig. 3. From the mathematical standpoint they complicate the situation exceedingly. From the physical standpoint they cause the secondary circuit to have a periodically-changing

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natural frequency and hence a bad wave form. In order to get a fairly smooth secondary wave, L_2 should be large compared with L_t—as large as the rest of the design will allow. Incidentally this large secondary inductance reduces the damping constant also, and improves the wave form due to somewhat suppressed attenuation, as will be seen later. In order to gain a first approximation to our solution, let us neglect L_t in the secondary for the moment and assume the voltage of Fig. 3 impressed upon a circuit of constant R, L, and C. After having familiarized ourselves with this simpler problem, we can then introduce the discontinuous secondary inductance and discuss its general effect.

So far we have said nothing at all about method of attack. The first of these, which suggests itself immediately upon glancing at the voltage of Fig. 3, is to represent this voltage by a Fourier series and operate with it as the impressed electromotive force upon the secondary oscillatory circuit with R_2, L_2, and C_2. Viewed from this angle, the method of frequency multiplication becomes one of filtering out harmonics contained in the distorted voltage wave.

There is, however, another very logical method of attack which suggests itself from the analogy between an electrical oscillatory circuit and the mechanical pendulum. If we were to subject an ordinary pendulum to periodically repeated pushes of short duration, we could obviously keep the system in motion in spite of its attenuation resulting from frictional forces. If the direction and frequency of the pushes or shocks were judiciously chosen with regard to the pendulum's own natural period, then the resulting amplitude in the steady state could reach large values. If, on the other hand, the frequency and direction of the repeated shocks were injudiciously chosen, the result would be one of misdirected effort and the amplitude attainable would be correspondingly small.

In an electrical oscillatory circuit a suddenly applied force produces damped oscillations. The same are produced by a suddenly relinquished force. We call these *transient effects.* If we repeat the same applications periodically we shall eventually arrive at a steady-state phenomenon which consists of a series of periodically recurring transients. Viewed from this angle, the present problem of frequency multiplication appears to be due to *shock excitation,* which is probably the term more commonly used to designate this method.

Having thus oriented the reader in our problem, we shall proceed to sketch the mathematics involved in the two methods of attack given above.

Using the notation illustrated in Fig. 3, we assume for the impressed voltage the following Fourier series in complex form:

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$$e(t) = \sum_{-\infty}^{+\infty} {}_k\, a_k e^{jk\omega t}, \tag{1}$$

in which $e(t)$ is the instantaneous secondary voltage at time t. Taking the origin of time as indicated in the figure, we find according to Fourier:

$$a_k = \frac{2jE \sin k\epsilon \frac{\pi}{2}}{\pi k} \cdot (-1)^{(k+1)/2} \tag{2}$$

with k restricted to odd integer values. We shall call ϵ the *breadth factor*. It varies between zero and unity as the breadth of the shock varies between zero and a half-period of the fundamental. Putting $\epsilon = 1$ in (2) we get:

$$(a_k)_{\epsilon=1} = \frac{-2jE}{\pi k},$$

which is the well-known coefficient for the square wave. The differential equation for the secondary circuit is

$$L_2 \frac{di}{dt} + R_2 i + \frac{1}{C_2} \int i dt = e(t). \tag{3}$$

The steady-state solution (particular integral) must be of the form

$$i = \sum_{-\infty}^{+\infty} {}_k\, b_k e^{jk\omega t}. \tag{4}$$

Substituting (1), (2), and (4) into (3) and equating coefficients of like frequency, we see that

$$b_k = \frac{a_k}{\Omega_k} \tag{5}$$

where

$$\Omega_k = R_2 + j\left\{ L_2 k\omega - \frac{1}{C_2 k\omega} \right\}. \tag{6}$$

The Fourier solution for this simple case is thus complete. Equation (6) we recognize as being a resonance denominator. Resonance to a given harmonic occurs when

$$L_2 k\omega = \frac{1}{C_2 k\omega}.$$

If we define

$$q_{20}^2 = \frac{1}{L_2 C_2} \tag{7}$$

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as the squared undamped secondary angular velocity, and

$$n_0 = \frac{q_{20}}{\omega} \tag{8}$$

as the undamped frequency multiplication, then the resonance condition obviously becomes

$$n_0 = k. \tag{9}$$

Hence we should get a maximum response in the secondary whenever it is tuned to an odd multiple of the primary, the resonance amplitude being given by

$$\frac{a_k}{R_2}.$$

For practical comparisons it is necessary to find the effective value of the secondary current. We have by the usual definition:

$$i_{eff}^2 = \frac{\omega}{2\pi} \int_0^{2\pi/\omega} i^2 dt. \tag{10}$$

In forming the square of the current we shall use (4) twice, with different summation indices, to avoid confusion. We get

$$i_{eff}^2 = \frac{\omega}{2\pi} \int_0^{2\pi/\omega} \sum_{-\infty}^{+\infty}{}_k\, b_k e^{jk\omega t} \cdot \sum_{-\infty}^{+\infty}{}_s\, b_s e^{jk\omega t} dt$$

$$= \frac{\omega}{2\pi} \sum_{-\infty}^{+\infty}{}_{ks}\, b_k b_s \int_0^{2\pi/\omega} e^{j(k+s)\omega t} dt.$$

Here all integrals vanish except those for which $k = -s$. Hence we get

$$i_{eff}^2 = \sum_{-\infty}^{+\infty}{}_k\, b_k b_{-k}. \tag{11}$$

But b_{-k} is the conjugate of b_k so that

$$b_k b_{-k} = |b_k|^2,$$

and thus

$$i_{eff}^2 = 2 \sum_{1}^{\infty}{}_k\, |b_k|^2. \tag{12}$$

But from the preceding:

$$|b_k|^2 = \frac{4E^2 \sin^2 k\epsilon \frac{\pi}{2}}{\pi^2 k^2 (R_2^2 + X_k^2)} \tag{13}$$

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with the substitution

$$X_k = L_2 k\omega - \frac{1}{C_2 k\omega} \cdot \tag{14}$$

Hence

$$i_{eff}^2 = \frac{8E^2}{\pi^2} \sum_{1,3,5\ldots}^{\infty}{}_k \frac{\sin^2 k\epsilon\frac{\pi}{2}}{k^2(R_2^2 + X_k^2)} \cdot \tag{15}$$

The above summation can be carried out and the expression for the effective current put into a closed form which is more useful for purposes of calculation than the infinite sum. We shall close our discussion of the Fourier method of attack with this brief outline and pass on to the more interesting shock-excitation method.

This method we shall also merely sketch briefly here. It is discussed in detail in the paper already cited. Let us review the simple transient problem involved. We know that the damped oscillations which take place in a circuit with R_2, L_2, C_2 in the absence of driving forces may be represented analytically by the form:

$$i_b = Im\{Be^{mt}\} \tag{16}$$

where Im denotes that the imaginary portion be taken, B is a complex amplitude which takes care of phase relations, and

$$\left.\begin{aligned} m &= -\alpha + jq_2 \\ \alpha &= \frac{R_2}{2L_2} \\ q_2 &= \sqrt{q_0^2 - \alpha^2} \\ q_0^2 &= \frac{1}{L_2 C_2} \end{aligned}\right\} \tag{17}$$

Equation (16) includes the rest condition for which $B = 0$. If a constant emf is impressed, or more generally, if a constant impressed force is suddenly changed by an amount ΔE, then the form of the oscillation is unaffected—its amplitude merely changes in magnitude and phase. The frequency and rate of damping must of necessity remain unaltered.

Suppose now that our circuit is oscillating according to (16) before a certain instant which we choose to call $t = 0$. At $t = 0$ we suddenly change the impressed force by ΔE. After $t = 0$, the oscillation will be given by

$$i_a = Im\{Ae^{mt}\}. \tag{18}$$

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At the instant $t=0$ we have to satisfy two conditions. First, the current must be continuous. This gives us

$$Im\{B\} = Im\{A\},$$

or if we write

$$\left.\begin{aligned} B &= B_1 + jB_2 \\ A &= A_1 + jA_2 \end{aligned}\right\} \tag{19}$$

we get

$$A_{20} = B_{20} \tag{20}$$

the subscript zero relating to the instant $t=0$.

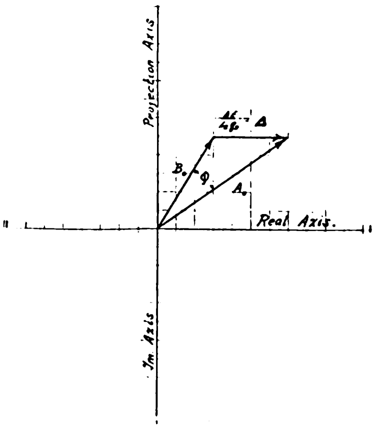

Fig. 4—Vector Relations in the Gaussian Plane for the Shock-Excitation Method of Attack.

Secondly, the discontinuity in the induced voltage must equal ΔE, i.e.

$$\left(\frac{di_a}{dt}\right)_{t=0} - \left(\frac{di_b}{dt}\right)_{t=0} = \frac{\Delta E}{L_2}.$$

Substitution gives

$$A_{10} = B_{10} + \frac{\Delta E}{L_2 q_2}. \tag{21}$$

Combining (20) and (21) we have

$$A_0 = B_0 + \frac{\Delta E}{L_2 q_2}. \tag{22}$$

Hence the amplitude after the disturbance equals the amplitude before the disturbance plus a real quantity which is proportional to the magnitude and direction of the sudden change in the constant applied force. Equation (22) is illustrated in the Gaussian plane by Fig. 4.

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The instantaneous value of current equals the projection of these rotating and decaying vectors upon the vertical axis. At the instant ΔE occurs the current has some value. It cannot change instantly, and hence the instantaneous change in the vector amplitude can be only in a horizontal direction as shown. This will give rise to an instantaneous phase shift ϕ, which will be zero only when B_0 is in a

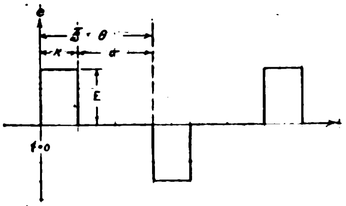

Fig. 5—Induced Transformer Voltage with Time Origin Chosen to Fit Shock-Excitation Method.

horizontal position. It is easy to see also that this phase shift may be in either direction, depending upon the direction (algebraic sign) of ΔE. Also A_0 need not always be larger than B_0 as a result of a suddenly applied (positive ΔE) emf. The amplitude can equally well be increased by suddenly removing an emf provided this is done at the proper instant. We shall let the reader illustrate these points for himself.

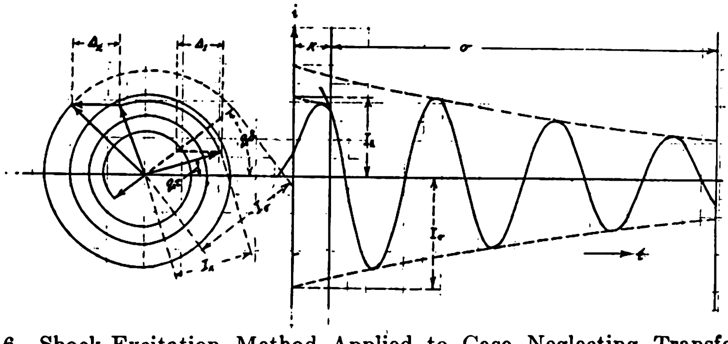

Fig. 6—Shock-Excitation Method Applied to Case Neglecting Transformer Inductance.
ϵ is chosen equal to 0.075; frequency multiplication equal to 7.0.

Our problem of handling the applied voltage of Fig. 3 for the oscillatory circuit R_2, L_2, C_2 is now very simple. During each fundamental period we have four sudden changes in voltage to consider, viz: The voltage jumps from zero to a constant value E and remains there for the duration of the shock; at the end of the shock duration it suddenly drops to zero and remains there until the advent of the negative shock; at this time it suddenly jumps to the constant negative value $-E$; and finally at the end of the negative shock it again drops back to zero

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and remains there until the beginning of the next fundamental period when the same sequence recurs.

We shall determine those amplitudes in magnitude and phase which obtain in the steady state. Here obviously the negative fundamental half-period must be an exact negative duplicate of the preceding positive fundamental half-period. Hence only one fundamental half-period need be considered. This half-period we shall divide into two parts. The duration of the shock we shall call the "coupling-period," because actually during this time the primary and secondary meshes are coupled with each other by the transformer mutual inductance. The remainder

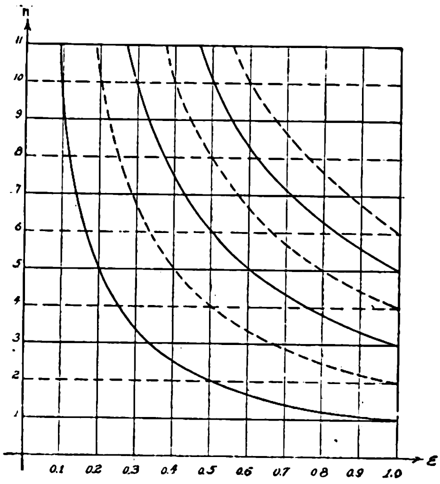

Fig. 7—Theoretical Relation between n and ϵ for Case of Neglected Transformer Inductance.
Intersections of full lines give rise to maxima, those of dotted lines to minima.

of the half-period we shall call the "period of free oscillation," because here the secondary circuit is virtually free to oscillate by itself due to the fact that the transformer is above saturation and its mutual inductance zero (according to our idealization of the *B-H* curve).

The steady state will be a succession of damped sinusoids whose amplitudes and phases change discontinuously at the boundaries of the coupling period and the period of free oscillation. Let us take the beginning of a positive shock as a time reference, i.e., as a reference for amplitudes and phases of the secondary current. Then if we denote the coupling period by the subscript k, and the period of free oscillation by the subscript σ, the periodicity of the steady state demands that at the beginning of the shock

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$$-I_\sigma e^{m\theta}+\frac{E}{L_2q_2}=I_k, \tag{23}$$

and at the end of the shock

$$I_k e^{mk}-\frac{E}{L_2q_2}=I_\sigma e^{mk} \tag{24}$$

where the quantities are as defined by (17) or illustrated by Fig. 5. This pair of equations is easily solved for the amplitudes giving:

$$\left.\begin{aligned} I_k &= \frac{E}{L_2q_2}\,\frac{1+e^{m\sigma}}{1+e^{m\theta}} \\ I_\sigma &= \frac{E}{L_2q_2}\,\frac{1-e^{-mk}}{1+e^{m\theta}} \end{aligned}\right\} \tag{25}$$

These amplitudes are of course complex and can be written

$$\left.\begin{aligned} I_k &= |I_k|\, e^{jq_2\tau} \\ I_\sigma &= |I_\sigma|\, e^{jq_2\delta} \end{aligned}\right\} \tag{26}$$

where τ and δ are the time phases.

The pair of equations (25) completely solve the case for the type of voltage illustrated by Fig. 5 applied to an oscillatory circuit with constants R_2, L_2, and C_2. Fig. 6 serves to illustrate the application of these equations to an arbitrarily chosen case with a breadth factor $\epsilon=\dfrac{k}{\theta}=0.075$ and a frequency multiplication $n=\dfrac{q_2}{\omega}=7$. The scale is immaterial since we are not interested in quantitative values here. The quantities Δ_1 and Δ_2 are the sudden changes in current amplitude. At the left we show the diagram of rotating and decaying vectors; at the right, the corresponding time plot.

It is interesting to note that the sudden drop in voltage is as effective in increasing the amplitude as the sudden rise, and that the discontinuities in the applied voltage alone produce an increase in amplitude. Whether the discontinuities produce amplitude changes (Δ_1 and Δ_2) which increase or decrease the absolute magnitude of the current depends upon the phases $q_2\tau$ and $q_2\delta$ as can readily be seen from the figure. These phases naturally depend upon the values of n and ϵ, so that it is evident that the question of a favorable or unfavorable result will depend upon these two parameters. Fig. 7 illustrates this dependence.[2] The full lines indicate favorable and the dotted lines unfavorable conditions. The combinations of n and ϵ, corresponding

[2] For the derivation of the equations involved here and for further illustrative examples we refer the reader to the reference given above from which the figures given here are taken.

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to the intersection of two full lines, give rise to maximum amplitudes, and those corresponding to the intersection of two dotted lines give rise

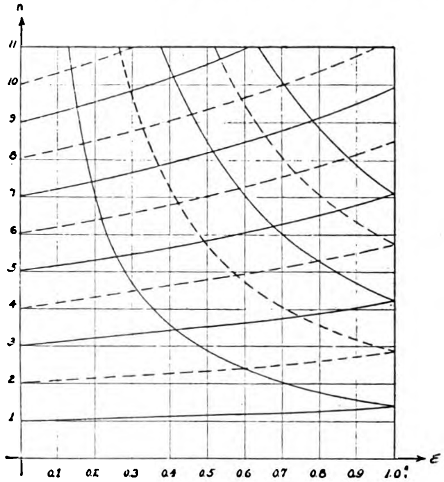

Fig. 8—Theoretical Relation between n and ϵ Taking into Consideration Transformer Inductance.
Full and dotted lines have same significance as for Fig. 7.

to minimum amplitudes. If it were possible to show a third (vertical) co-ordinate for secondary r.m.s. current, then we should have a relief

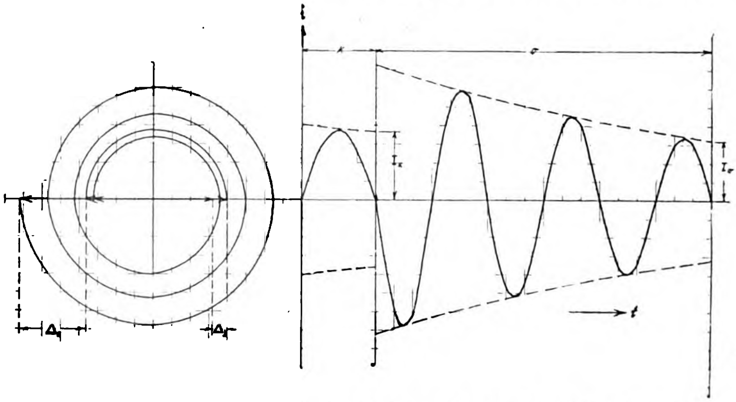

Fig. 9—Shock-Excitation Method Applied to Case in Which Transformer Inductance is Considered.
Condition for maximum secondary response with $n = 7.414$ and $\epsilon = 0.19$.

map, the peaks of which would correspond to current maxima, and the hollows to current minima. The intersections of a dotted with a full line would correspond to a saddle point or "ridge." Cross-sections

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of this relief map would show the current variation for constant n and varying ϵ or constant ϵ and varying n.

An investigation of such cross-sections is interesting from the following point of view. If we go back to the standpoint of filtering harmonics then we should expect maxima to occur whenever the secondary is tuned to an odd multiple of the primary, since the Fourier analysis of the impressed voltage contains only odd harmonics. However, although this is true, we are very likely to overlook the importance of the effect of ϵ upon the magnitude of the Fourier coefficients. For instance, with $\epsilon = 2/3$ and $n = 3$ we should get a very small secondary current. With a given circuit and fixed shock duration, therefore, we should expect to find that certain odd harmonics in the secondary are very much suppressed as a result of an unfavorable ϵ.

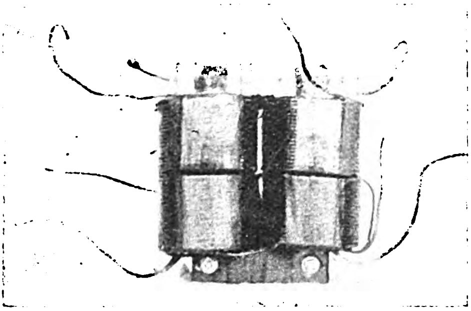

Fig. 10—Transformer Used in Experimental Work.

The picture of the actual state of affairs is still incomplete since we have thus far overlooked the fact that the total secondary inductance is *not constant.* For the shock duration it is equal to the external secondary inductance plus the secondary transformer inductance, and for the remainder of the fundamental half-period it is equal to the external secondary inductance alone. This additional feature alters the situation quite considerably and gives rise to the anomalies which the present experimental investigation proposed to illustrate.

The method of treating the problem from the standpoint of recurring transients is still applicable to the actual problem with non-constant secondary inductance, but involves—in addition to the transient set up by the sudden voltage change—also a transient due to the simultaneous inductance change. The effects of these transients superpose to form the net effect.

The application of the result is the same as for the simple case just treated, and in vector form has the same appearance as shown by Fig. 4.

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The only difference is that the sudden amplitude change Δ depends not only upon ΔE but also upon the change of inductance.[3] Also the magnitude of Δ will not be the same at the beginning as at the end of the shock. We shall not take the time to go into details here, but prefer to show the theoretical results and compare them with the experimental results found in this investigation.

Fig. 11—Primary Current and Transformer Induced Voltage with 110 volts, 60 cycles impressed. Primary current 2 amperes.

Here again the parameters n and ϵ govern the occurrence of maxima and minima. The plot which corresponds to Fig. 7 for the simple case is given in Fig. 8. Comparing Figs. 7 and 8 we see that now the maxima do not occur for odd values of n, but for slightly larger values depending upon ϵ. Why this should be the case can be seen from Fig. 9, which illustrates the application to a case where maxima conditions are fulfilled. Here we see the effect of the higher inductance during the shock very clearly, in that the half-period there is considerably

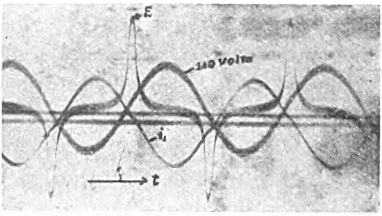

Fig. 12—Primary Current and Transformer Induced Voltage with 110 volts, 60 cycles impressed. Primary current 7 amperes.

longer than for the rest of the fundamental half-period. Also the amplitude is smaller, as it should be, considering that the same inductive countervoltage is here produced with a larger total inductance. Now, since the frequency multiplication n is taken as the ratio of the frequency during the period of free oscillation to the fundamental frequency, it is quite clear that this ratio must now be somewhat larger than an odd number of half cycles. This is evident from Fig. 9 and needs no further comment.

[3] For the derivation of this case see the reference above.

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As a matter of fact it becomes rather difficult to define the frequency multiplication accurately, since there are actually three different ways in which this may be done. We could take the ratio of the frequency either during shock or during the remaining fundamental half-period to the fundamental frequency, or an average of these two values as that figure which is to define the frequency multiplication of the circuit as a whole. However, since the time of free oscillation greatly predominates over the duration of shock, and since in most practical circuits the secondary inductance of the transformer is small compared with the external inductance, it seems most reasonable to call the free oscillation frequency the secondary frequency and base the frequency multiplication upon it.

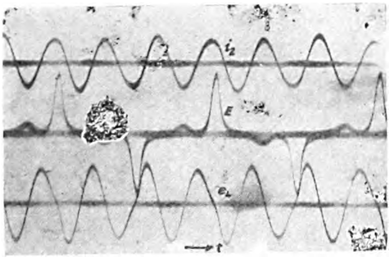

Fig. 13—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Third Harmonic.
Primary current 5.5 amperes, secondary current 2.35 amperes. Condition of oscillations of the "second kind." Efficiency 51.2 per cent.

With this brief review of the fundamental principles involved, we are prepared to consider some of the experimental results obtained and discuss them in the light of the theory.

The circuit used was exactly that shown in Fig. 1. A 60-cycle alternator was used as a primary source and the primary inductance was of the iron-core type, but so dimensioned that its core remained well below saturation throughout the cycle. The transformer, shown in Fig. 10, was of the closed magnetic circuit type, and provided with four windings with taps so that various ratios could be tried out. To all of the following figures and discussions a 1 to 1 ratio applies.

Since it was desired to vary the duration of shock or the parameter ϵ, the transformer was first provided with a variable air gap for this purpose. This method proved very ineffective, however, and ϵ was therefore varied by means of the primary current amplitude alone. That this is possible may easily be realized from the fact that with a small primary amplitude the transformer core remains below saturation for a longer time that it does for a large primary amplitude. This

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is illustrated by the oscillogram Figs. 11 and 12, which were taken with primary currents of two and seven amperes respectively. This method of varying the shock duration had the disadvantage of varying the induced voltage maximum at the same time. This fact could, however, easily be taken into account in interpreting the results, so that the method was still perfectly suitable for comparative work.

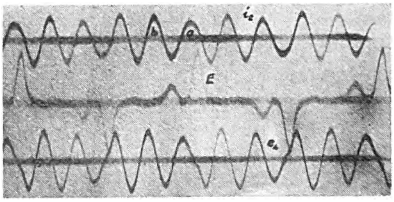

Fig. 14—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Fifth Harmonic.
Primary current 5.0 amperes, secondary current 1.47 amperes. Condition of oscillations of the "second kind."

How the secondary period is lengthened during shock, by reason of the higher total secondary inductance, is shown by oscillogram Figs. 13, 14, and 15. Fig. 13 is taken with the secondary tuned to the third harmonic. At the top is the secondary current, in the center the secondary terminal voltage, and at the bottom the voltage induced in the secondary external inductance. Here we see that the half-period

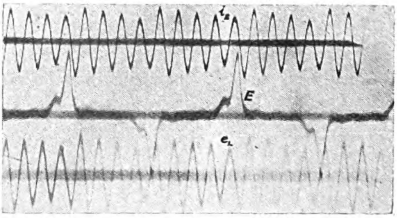

Fig. 15—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Ninth Harmonic.
Primary current 8 amperes, secondary current 0.8 ampere. Condition of oscillations of the "second kind."

corresponding to the shock duration (total secondary inductance equal to transformer secondary plus external) is appreciably longer than the following two periods which correspond to free oscillation of the secondary (total secondary inductance equal to external only). The same effect is illustrated for the fifth harmonic in Fig. 14 and for the ninth harmonic in Fig. 15. We shall refer again to these figures in connection with secondary voltage distortion.

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In Fig. 9 we have shown a theoretical condition for maximum secondary response. This condition requires that the secondary half-period during shock should equal the duration of the shock, and that the remaining fundamental half-period should be just long enough to accommodate an even number of half-periods of free oscillation. In Fig. 16 we show an oscillogram which was taken with these conditions fulfilled for the seventh harmonic. The shock duration is just right, the result is smooth and regular, and the efficiency and secondary current are maxima. Incidentally the smoothness of the current can best be recognized by an inspection of the voltage induced on the secondary external inductance, since this voltage magnifies any irregularities that may be present in the current.

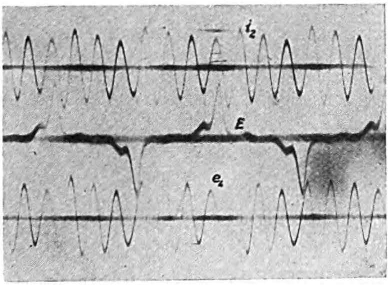

Fig. 16—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Seventh Harmonic.
Primary current 6 amperes, secondary current 0.88 ampere. Condition of oscillations of the "second kind," and shock duration adjusted for smoothest wave form. Note this in the inductance voltage.

What happens when the shock duration is too short is illustrated by Fig. 17, which was taken with the secondary tuned to the third harmonic. A shock duration which is too short need not necessarily give rise to these irregularities, however, as may be seen in Fig. 13, which is taken for exactly the same primary conditions and hence for the same shock duration on the basis of no-load conditions. These features are intimately tied up with the question of oscillations of the *first and second kind.*

In order to discuss this point adequately it is necessary to recall some well-known facts about frequency multiplication by means of the a.c. arc. The frequency multiplication circuit for the arc is identical with that shown in Fig. 1 except that the transformer is replaced by an arc. We know from our arc circuit theory that the voltage produced across the arc by an alternating current flowing through it is substantially a square wave. This square-wave voltage is then impressed

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upon the secondary oscillatory circuit which responds to the harmonics contained in it. This is the same situation as for the case of frequency multiplication by inductive means with the special condition of a breadth factor of unity or a shock duration of half a fundamental period, assuming of course that a breadth factor of unity were pos-

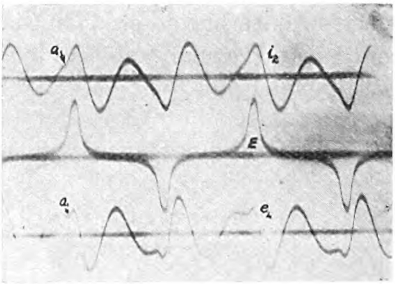

Fig. 17—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Third Harmonic.
Primary current 5.5 amperes, secondary current 0.54 ampere. Condition of oscillations of the "first kind" and shock duration too short for smoothest wave form. Note irregularities in inductance voltage. Efficiency 28 per cent. Compare with Fig. 13 for same primary conditions but operation within region of oscillations of the "second kind."

sible with the transformer method. The Fourier treatment is identical for these two problems.

From this similarity it is to be expected that other peculiarities found in the arc circuit should be recognizable in the magnetic circuit

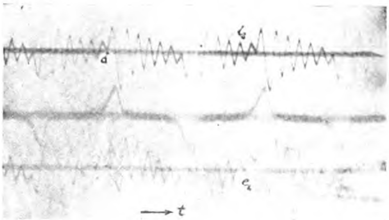

Fig. 18—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Twelfth Harmonic.
Primary current 3 amperes, secondary current too small to read. Phase reversal at point *a*. Phase reversals must occur when secondary is tuned to even harmonics.

also. Now we know that in the arc circuit we may obtain oscillations in the secondary for which the arc current reverses only once for each fundamental half-period. Such oscillations are known as oscillations of the *first kind*. If the secondary current amplitude exceeds a certain

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limit, then the arc current reverses *twice* in each fundamental half-period and the result is called oscillations of the *second kind.* If the arc current reverses *three times,* we get oscillations of the *third kind,* etc.

An exactly analogous phenomenon takes place in the magnetic circuit discussed in this paper. Instead of *arc current,* however, we have to consider the *transformer flux.* If n times the secondary current amplitude is less than or equal to the primary current amplitude (n being the frequency multiplication and the transformer ratio unity), then the flux in the transformer core reverses only once for each fundamental half-period and the resulting secondary oscillations should be called oscillations of the first kind. If the secondary amplitude exceeds this value, then the transformer flux reverses twice in each fundamental half-period, and the result should be termed oscillations of the *second kind* by analogy to the arc circuit conditions.

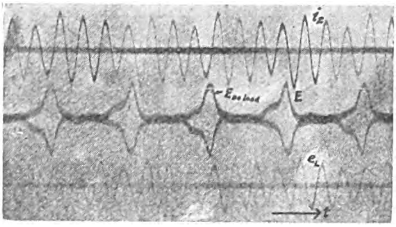

Fig. 19—Secondary Current, Transformer Voltage, and Secondary Inductance Voltage for Seventh Harmonic.
Primary current 3 amperes, secondary current 0.46 ampere. Oscillations just within region of "second kind." Transformer voltage shown under load and no-load conditions to emphasize distortion.

Without going into more details we can now discuss the conditions shown in Figs. 13 and 17. Both are for a frequency multiplication of three and a primary current of 5.5 amperes. Fig. 17 is taken for a secondary current of 0.54 ampere, while Fig. 13 is taken for a secondary current of 2.35 amperes. Hence Fig. 13 represents oscillations of the second kind, and Fig. 17 oscillations of the first kind. This can also be seen directly by examining the transformer voltage. In Fig. 17 it is smooth except for the peak, while in Fig. 13 it contains an additional hump. This additional hump means an additional reversal of the transformer flux, and hence oscillations of the second kind. Oscillations of the second kind occur for low secondary resistance and low attenuation. The effect produced is to broaden the voltage shock at the base and thus avoid, to a certain extent, the appearance of secondary current irregularities. Unless the shock duration is

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properly adjusted, therefore, it is desirable from the standpoint of smoothness to operate in the region of oscillations of the second kind. We shall later show that maximum efficiency is also obtained at the point where oscillations of the second kind begin, so that this should be the operating condition sought after. Figs. 14 and 15 also show conditions where oscillations of the second kind were obtained.

We have one other interesting point to illustrate by oscillograms, and that is in connection with the production of even harmonics in

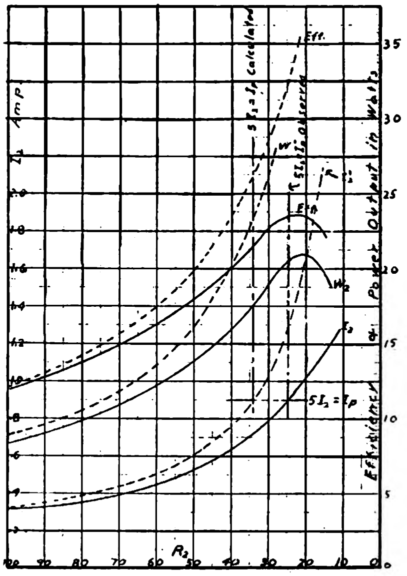

Fig. 20—$N=5$. Comparison of Observed and Calculated Results.
——————— Observed
- - - - - - - Calculated
$I_p=4.5$ amperes

the secondary. In order to be able to recognize them as such, it will be necessary to recall the following: it is quite clear that the secondary current during a negative fundamental half-period must be an exact negative duplicate of the preceding positive fundamental half-period, since there is no reason for positive and negative dissymmetry. Hence it follows that in each fundamental half-period there be *exactly* an odd number of secondary half-periods—including the sudden phase shifts as illustrated by ϕ in Fig. 4. If, therefore, we tune the secondary to an even multiple of the primary, then the sum of the sudden phase shifts during any fundamental half-period must necessarily equal π radians or be equivalent to one secondary half-period. It is this phase shift which makes up the discrepancy between the number of secon-

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dary half-periods actually accomplished by the current and the necessary odd number demanded by the symmetry of the steady state.

If the shock duration is adjusted so that it is equal to an odd number of secondary half-periods below saturation, and the secondary is tuned to an even multiple of the fundamental, then this phase shift of π radians will occur at *one point* and result in a *phase reversal* of the secondary current. In the oscillogram Fig. 18 we show such a condition for the secondary tuned to the twelfth harmonic. The point

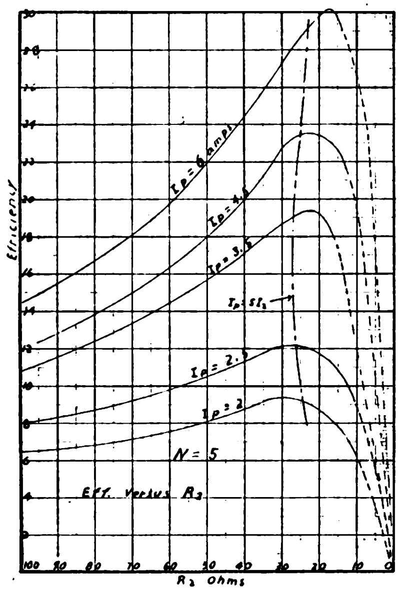

Fig. 21—Efficiency vs. Load Resistance for Various Primary Currents with $N = 5$.

of reversal is marked at *a*. The shock duration is approximately equal to one-quarter of the fundamental half-period.

The question of transformer secondary terminal voltage distortion has already been discussed to some extent in connection with oscillations of the second kind. The oscillogram, Fig. 19, serves, however, to illustrate this still more clearly. Here we have a frequency multiplication of seven and a condition of operation which is just within the region of oscillations of the second kind. Both the transformer no-load and load voltage are shown. They were taken in opposite phase positions so that the picture would be clear and easy to read. The effect of the secondary current is to broaden the base of the shock and make it more pointed. The reason for this can easily be shown by drawing in the flux curve by means of the magnetization curve.

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A large amount of data were also taken on power output and efficiency vs. frequency multiplication, primary, and secondary current. A few of the more interesting curves will be given here. Fig. 20, for instance, shows the secondary current, power output, and efficiency as functions of the secondary load resistance for a frequency multiplication of five and a primary current of 4.5 amperes. The broken line is drawn in at the dividing line between oscillations of the first and second kind. It is interesting to note that maximum efficiency coincides with maximum power output and that these maxima occur just within the region of oscillations of the second kind. This latter point is fortunate, since we have already seen that this operating condition also gives us the smoothest secondary wave form.

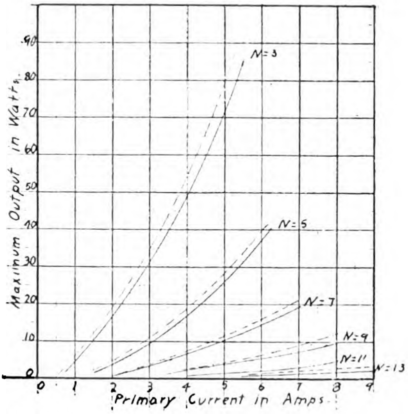

Fig. 22—Curves of Maximum Output vs. Primary Current for Constant Ratios of Frequency Multiplication.
———————— with 120 turns on secondary
- - - - - - - - with 160 turns on secondary

Fig. 21 shows the efficiency as a function of the load resistance for various values of primary current and a multiplication of five. Again the broken line indicates the division between oscillations of the first and second kind. Here it is interesting to note that the maximum efficiency increases with increasing primary current. How long this would continue we were unfortunately unable to determine, since it was not possible to go to higher primary currents without danger of primary condenser and inductance insulation breakdown.

Figs. 22 and 23 show the maximum power output and efficiency as functions of the primary current for various values of frequency multiplication. They show quite conclusively that this method rapidly becomes very inefficient for high values of multiplication. Thus it is clear that a multiplication of nine by two successive stages of three

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would be considerably more efficient than by one single stage. This question of multistage frequency multiplication will be discussed more thoroughly in a following paper on multiphase and multistage methods of extending the present fundamental principle.[4]

In conclusion we may state that this method of frequency multiplication is useful only where perfect wave form is not essential. The wave form is improved by making the external secondary inductance large compared with that of the transformer, and by operating just within the region of oscillations of the second kind as defined in this paper. Since the efficiency increases with increased transformer inductance, it will be a matter of judgment as to how far to go in the relative magnitude of external to transformer inductance. In the

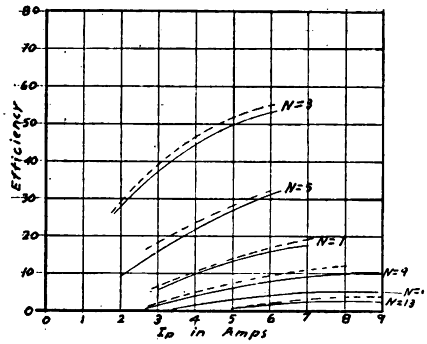

Fig. 23—Curves of Maximum Efficiency vs. Primary Current.
———————— with 120 turns on secondary
– – – – – – – with 160 turns on secondary

theoretical Fig. 9 this ratio was taken as unity, and this figure thus gives us a good idea of what such a ratio does to distort the secondary current. In the cases for which the oscillogram figures are here given the ratio of external inductance to total was about 0.7. The resulting distortion on this account is certainly not excessive. For further data as well as many interesting oscillograms and figures relative to this method of frequency multiplication we refer the reader to the complete report[5] of this investigation, which is filed in the library of the Massachusetts Institute of Technology, and of which this paper is merely an abstract.

[4] To appear in a forthcoming issue of the PROCEEDINGS.

[5] P. T. Rumsey, "Experimental Investigation of Frequency Multiplication by Means of Iron-Core Coupled Circuits." M.I.T. thesis.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

ON THE SHORT-WAVE LIMIT OF MAGNETRON OSCILLATIONS*

By
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Summary—*In this paper the theoretical considerations and experimental results pertaining to the short-wave limit of magnetron oscillations are given.*

Oscillations of only a few centimeters wavelength were successfully produced.

INTRODUCTION

FOR several years research work on the generation of very short waves has been carried on by Professor Yagi and others in our laboratory. While engaged in this work, the writer found it possible to produce very short wavelength oscillations with magnetrons of special form.

Certain of the details of this work have been published.[1] Since then the experiments have been extended and oscillations can now be produced with increased intensity and with even shorter wavelengths. Only the method and prospect of obtaining shorter wavelengths are discussed in this paper. The details regarding the means of increasing the intensity of the oscillations will be dealt with in another paper.

1. THEORETICAL CONSIDERATIONS

The distribution of the space charge in the magnetron is quite similar to that in the triode when the latter is connected to produce oscillations similar to those observed by Barkhausen and Kurz. By considering a "virtual cathode" and a "virtual grid," and an equivalent anode potential, the existence of a negative resistance can be proved in the same way as has been done by Tonks[2] in the case of the Barkhausen and Kurz type of oscillations. Such a negative resistance characteristic gives rise to a state of instability. Sometimes oscillations may be due to a dynatron characteristic and at other times oscillations may be due to the dispersion or discharge of the space charge.

It seems very likely that oscillations in the magnetron occur in the following manner:

* Dewey decimal classification: R133. Original manuscript received by the Institute, October 5, 1928. Revised copy received, November 7, 1928.

[1] Hidetsugu Yagi, "Beam Transmission of Ultra Short Waves," PROC. I. R. E., **16**, 715; June, 1928.

[2] Lewi Tonks, *Phys. Rev.*, October, 1927.

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When the magnetic field is greater than the critical value, the anode current is cut off and there is an accumulation of charges near the anode and cathode. The accumulation of charges near the anode is the true seat of oscillatory phenomenon.

As soon as the conditions become such that a negative resistance is possible, the accumulated charges begin to be dispersed or discharged and the state of negative resistance disappears until a further accumulation of charges occurs. The time required for the accumulation of these charges after the preceding discharge has occurred is approximately equal to the time, t, which will be required by an electron in travelling from the cathode to the region near the anode where the charge density is greatest. This process of accumulation and discharge will be repeated and cause oscillations.

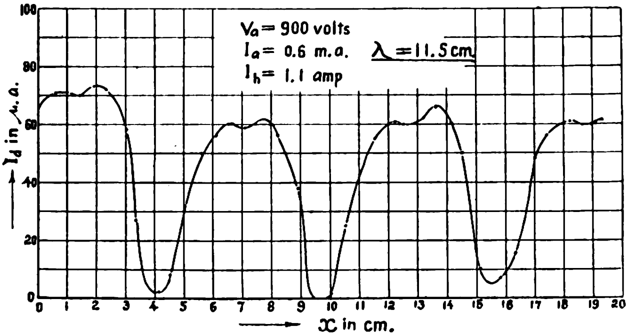

Fig. 1

The period T, which corresponds to one cycle of oscillation will be

$$T = t + t'$$

where t' represents the time required for the electronic discharge. We have no means at present for estimating the value of t'. The value t, however, may be calculated.

Since the phenomenon is alternating, it may not be unreasonable to assume that t' is approximately equal to t. Then the wavelength may be calculated by the following semi-theoretical formula,

$$\lambda_0 = 2ct \qquad (1)$$

where c represents the velocity of light.

The time t can be obtained from a graphical solution of the following equation:

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The measured wavelength and the values calculated from (12) are given in Table I, where the "difference" in the third column is the percentage difference as follows:

$$\text{"difference"} = \frac{\lambda_0'' - \lambda}{\lambda} 100 \text{ per cent.}$$

It may be shown that, under the condition of the maximum oscillation, the radius of the anode cylinder should be related definitely to the square root of the applied anode voltage. The maximum os-

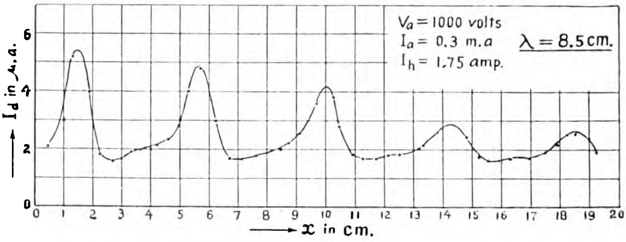

Fig. 2

cillation generally occurs at the neighborhood of the critical field strength and consequently, when H is given, the following relation should exist, which is deduced by putting $r' = r_a$ in equation (5).

$$\left. \begin{aligned} r_a &= \sqrt{\frac{8mc^2}{H^2 e} V_a} \\ \text{or} \quad &= \frac{6.7}{H}\sqrt{V_a} \text{ cm} \end{aligned} \right\} \quad (13)$$

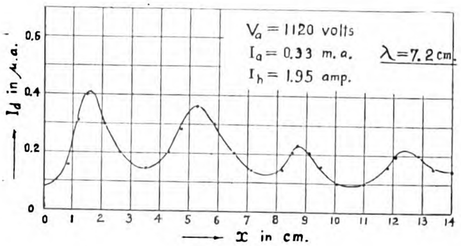

Fig. 3

All of the above equations are applicable to most of the cases of the magnetron oscillations, except the cases where the external circuits are important.

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2. Experimental Results

Experiments have been carried out with the small split-anode magnetrons as shown in Table II.

TABLE I

Wavelengths Obtained by Experiments (cm)	Wavelengths Calculated from Eq. (12) (cm)	Difference (%)	
38	41.2	+ 8.4	Table 4 in *Jour.* I.E.E. of Japan, No. 467
36	39.3	+ 9.2	
32.3	36.5	+13.3	
42	43.5	+ 3.6	Table V in the above paper.
38	41.3	+ 8.7	
31	35.3	+13.9	
28.5	32.4	+13.7	
52	52.3	+ 0.6	Table I in *Jour.* I.E.E. of Japan, No. 469.
50	49.1	− 1.8	
49	44.8	− 8.6	
39	39.4	+ 1	
36.5	36.3	− 0.6	
33	34.2	+ 3.6	
32	33.3	+ 4	
70	62	−11.4	Table II in the above paper.
51	52	+ 2	
44	44.8	+ 0.9	
39.5	40.8	+ 3.3	
37	36.1	− 2.4	
34	34.2	+ 0.6	
47	41.5	−11.7	Table 7 in the Technology Report of the Tohoku Imp. Univ. Vol. 7, No. 4.
42	42	0	
45	46.4	+ 3.1	
54	55.2	+ 2.2	
52	52.7	+ 1.4	
48	48.2	+ 0.4	
11.5	11.2	− 2.6	Figs. 1, 2, 3, 4, and 5 in the present paper.
8.5	7.1	−16.5	
7.2	6.4	−11.1	
6.8	5.7	−16.1	
5.6	5.4	− 3.6	

TABLE II

No. of Vacuum Tube	Diameter of Anode Cylinder (cm)	Length of Anode Cylinder (cm)	Diameter of Cathode Wire (cm)
I	0.5	0.8	0.01
II	0.35	0.5	0.01
III	0.25	0.5	0.01

The experimental results obtained using tube No. 2 are shown in Figs. 1, 2, 3, and 4. In these figures, V_a, I_a, and I_h denote anode voltage, the anode current, and the field current, respectively. In this case the field intensity H is 1050 gauss per ampere.

Throughout the measurements, the crystal detector bridging the parallel wires was kept at a fixed position and the short circuiting condenser was moved back and forth to obtain the wavelength

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$$\frac{dt}{dr}=\frac{1}{\sqrt{\frac{2eV_a}{m}\frac{\log\frac{r}{r_o}}{\log\frac{r_a}{r_o}}-\frac{H^2e^2r^2}{4m^2c^2}\left(1-\frac{r_o^2}{r^2}\right)^2}} \tag{2}$$

where t, r, r_o, r_a, V_a, and H represent the time, the distance from the central axis, the radius of the cathode, the radius of the anode cylinder, the anode voltage, and the intensity of the magnetic field, respectively. e and m are the charge and the mass of an electron. c denotes the velocity of light as in the case of (1).

For the sake of simplicity, let us now assume that r_o is extremely small; then (2) becomes the following:

$$\frac{dt}{dr}=\frac{1}{\sqrt{\frac{2eV_a}{m}-\frac{H^2e^2}{4m^2c^2}r^2}}\,. \tag{3}$$

The graphical solution of (3) shows that the assumption introduces only a small error, and this error is not serious since it only results in a calculated value of wavelength which is somewhat too small.

By integration of (3), we obtain

$$t \doteqdot \frac{2mc}{He}\sin^{-1}\frac{r'}{\sqrt{\frac{8V_amc^2}{H^2e}}} \tag{4}$$

where r' is the value of r which satisfies the condition,

$$\frac{dr}{dt}=0$$

or

$$r'=\sqrt{\frac{8V_amc^2}{H^2e}}\,. \tag{5}$$

From (4) and (5) we obtain

$$t \doteqdot \frac{\pi cm}{eH}\,. \tag{6}$$

Assuming on the other hand r_o extremely large, we have

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$$\frac{dt}{dr}=\frac{1}{\sqrt{\frac{2eV_a}{md}r-\frac{H^2e^2}{m^2c^2}r^2}} \tag{7}$$

in which d represents the normal distance between two electrodes which now form two parallel plates, and r represents the distance from the cathode surface.

The following equations (8), (9), and (10) correspond to the equations (4), (5), and (6).

$$t=\frac{2mc}{He}\sin^{-1}\sqrt{\frac{deH^2}{2V_amc^2}r'} \tag{8}$$

$$r'=\frac{2V_amc^2}{deH^2} \tag{9}$$

$$t=\frac{\pi cm}{eH} \tag{10}$$

Equations (6) and (10) are exactly the same, so that in both cases the wavelengths can be calculated by the following equation,

$$\lambda_0'=\frac{2\pi c^2m}{eH}$$

or

$$=\frac{10650}{H}\text{cm} \tag{11}$$

This equation shows that the wavelength depends only upon the intensity of the applied magnetic field, and is independent of the value of the applied anode voltage. These characteristic features are in exact conformity with what have been observed in actual experiments.[3]

The corresponding equation representing the experimental results has the form

$$\lambda_0''=\frac{13000}{H} \tag{12}$$

so that the only difference is in its numerical value.

The wavelengths calculated from (11) should naturally be smaller than those obtained from (2), and hence the proportional difference of about 20 per cent between (11) and (12) can be accounted for.

[3] Note: Fig. 28 of Yagi's paper, loc. cit., does not show the relation between the applied anode voltage and the wavelength with the meaning as here referred to. In that case the intensity of the magnetic field was always so adjusted that the oscillation becomes maximum at every setting of the applied anode voltage.

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measurements. I_d and x represent the detector current and distance through which the short circuiting condenser was shifted.

From the figures, it can be seen that the equations given do apply to the case where the wavelengths are extremely short. (Refer to Tables I and II and to equations (12) and (13)).

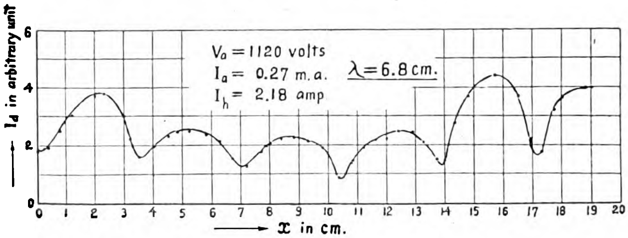

Fig. 4

The shortest wave that has ever been obtained by this method was produced with tube No. 3. The oscillations in this case had a wavelength of 5.6 cm. The potential distribution on the parallel wires is shown in Fig. 5.

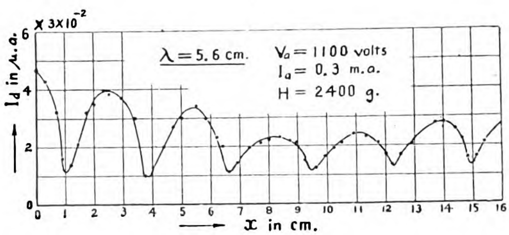

Fig. 5

3. Conclusion

The theory proposed in the present paper seems to be a very reasonable one, though not very exact, since there are many details left out of consideration. These details are to be studied further.

The author is of the opinion that this theory is also applicable to the type of oscillations observed by Barkhausen and Kurz.

The following is an approximate example of the limit of the wavelength, which can be obtained with a magnetron oscillator. These values are estimated from the summary of theoretical considerations given in an early part of the paper.

Suppose that the intensity of the applied magnetic field can be raised to 20,000 gauss—equation (12) gives the wavelength to be 6.5 millimeters. Then, if we desire to obtain maximum oscillation the

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anode radius must be made 0.34 millimeter, (13), and at the same time the anode potential must be raised to 10,000 volts. The practicability and even the possibility of such a small scale experiment is questionable. We may say, however, that the practical lower limit of the wavelength of oscillations produced by the magnetron under the condition of maximum oscillation is of the order of a few millimeters. If the intensity of oscillation be disregarded, it may be possible to reduce the wavelength to a slightly lower value.

The shortest wavelength actually produced so far (5.6 cm) is less than half of that previously reported,[1] and about one fourth of the lowest wavelength obtained by the Barkhausen and Kurz method. The intensity of the oscillations at this wavelength might be further increased if the anode voltage could be raised to the point indicated by (13).

The harmonic waves are not considered for the time being, though they are by no means unimportant, since the oscillations obtained from the magnetrons are apt to be distorted waves.

We can produce very short wavelength oscillations of two different types with a special form of magnetron. These two types were (a) the oscillation whose frequency is approximately independent of the external circuit, and (b) the one whose frequency depends principally on the external circuit. Most of the intense oscillations which we were able to produce were combinations of types (a) and (b).

The oscillations obtained by Professor Zacek, of which we learned recently, seem to correspond to type (a) just as do these described in the present paper.

The author wishes to express his thanks to Professor Yagi for his valuable instructions and suggestions.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

THE USE OF THE ELECTRON TUBE PEAK VOLTMETER FOR THE MEASUREMENT OF MODULATION*

By
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METHODS of measuring the degree of modulation of a radio-frequency generating set have been described[1] by Nelson and Conrad, but both systems have the disadvantage that they measure the amplitude of the audio-frequency current before it is impressed on the radio-frequency current. The oscillograph has been used extensively and provides an ideal method of measuring the degree of modulation, but it is an expensive instrument and not readily portable.

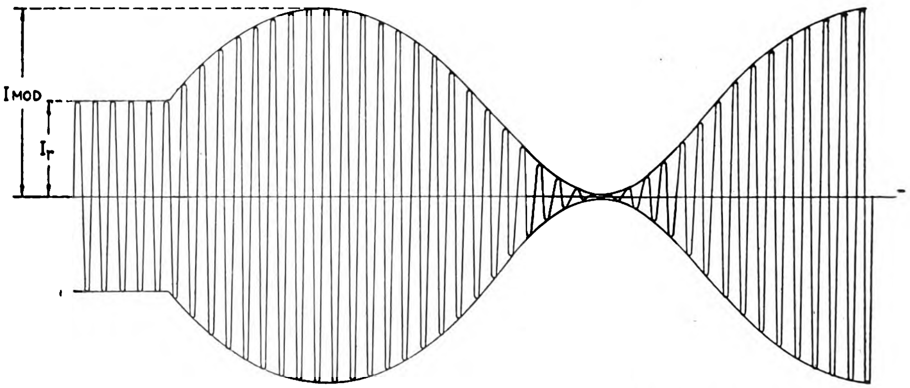

Fig. 1—Sine Modulated Radio-Frequency Current.

The electron tube peak voltmeter described by van der Bijl[2] may be used as a means of measuring the modulation of a radio-frequency current whose maximum value for complete modulation varies between the limits zero and two times the maximum value of the unmodulated radio-frequency current (see Fig. 1). Such a current is produced by the so-called "Heising" or "constant current" system of modulation. In brief, if the peak value of the radio-frequency current

* Dewey decimal classification: R275. Reprinted at the request of The Committee on Standardization from *Jour.* Opt. Soc. of Amer., 9, December, 1924. Published by permission of the Director of the Bureau of Standards of the U. S. Department of Commerce.

[1] Nelson, U. S. Patent 1,478,050. Conrad, U. S. Patent 1,477,316.

[2] van der Bijl. Thermionic Vacuum Tubes, p. 367.

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is measured without modulation and then the modulation is applied and the peak value again measured the ratio

$$\frac{(I_{MOD}-I_r)\times 100}{I_r} = \text{per cent modulation} \qquad (1)$$

I_{MOD} = peak value of the modulated current.

I_r = peak value of the unmodulated radio-frequency current.

The actual apparatus needed is shown in Fig. 2. The current I flows through the resistance R (1000Ω), so that measurements on the

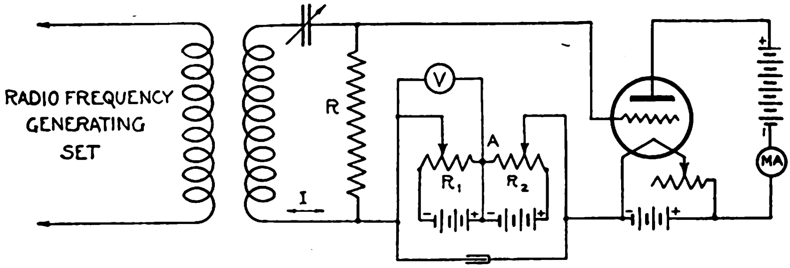

Fig. 2—Electron Tube Peak Voltmeter Used to Measure Percentage Modulation.

variation of the voltage across the fixed resistance R give values which are proportional to the variation in that current. The current I is proportional to the output current of the radio-frequency generating set.

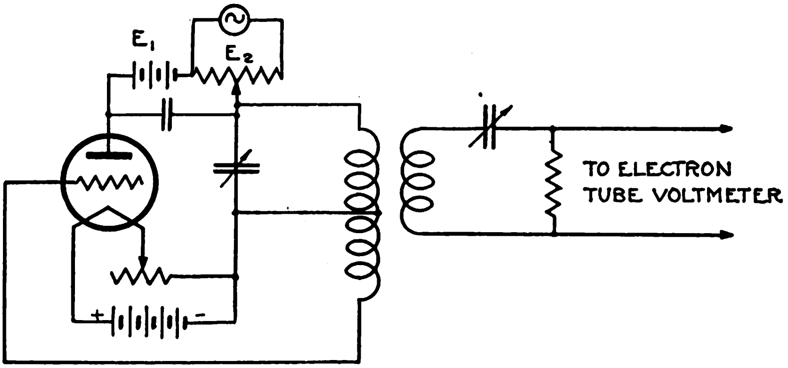

Fig. 3—Generating Circuit Used to Test Method.

The principle of the voltmeter is that the voltage of the grid is adjusted until the milliammeter just fails to show a plate current. The resistance $R_2(0-2000\Omega)$ with $R_1(0-100\Omega)$ set at zero (A) is adjusted until the plate milliammeter reads zero. The unmodulated radio frequency is then applied and R_1 adjusted until the milli-

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ammeter again reads zero. The voltmeter V gives then the peak value of the radio-frequency voltage which is proportional to the current I_r. The radio frequency is then modulated and R_1 further increased until the plate milliammeter again reads zero. The voltmeter will then read the peak value of the modulated radio-frequency voltage which is proportional to I_{MOD}. The percentage modulation then follows from (1).

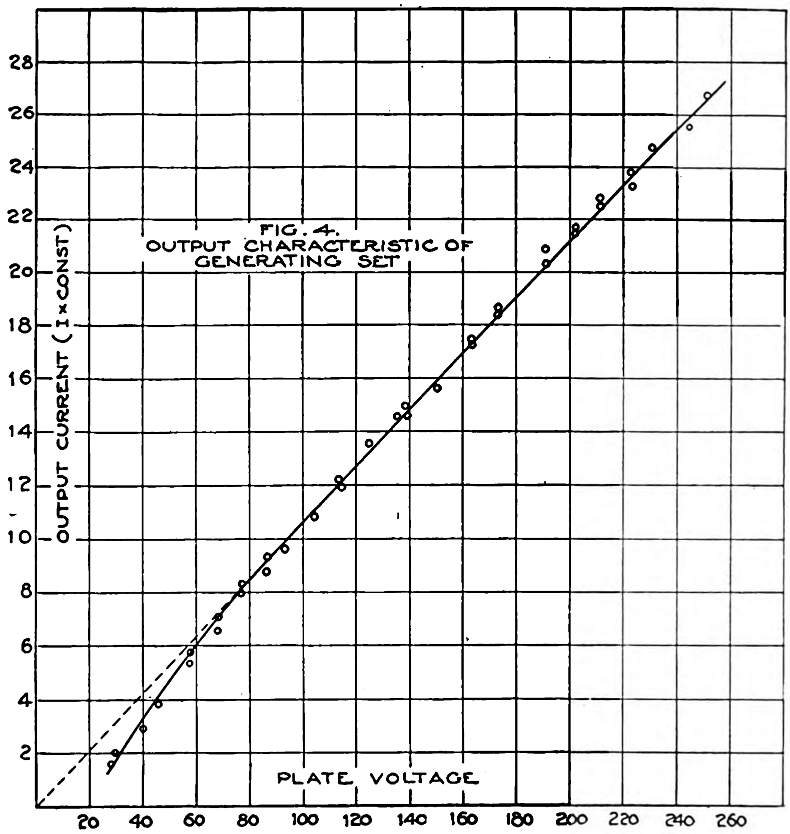

Fig. 4—Output Characteristic of Generating Set.

To test this method the generating circuit shown in Fig. 3 was used. The variation of output current with plate voltage for this circuit is shown in Fig. 4. Since, for the upper portion of the curve the output current is proportional to the plate voltage, the degree of upward modulation may be calculated from the measured values of E_1 and E_2. E_2 was either a 500 or 60-cycle sinusoidal a.c. source having an effective voltage of 120 volts. E_1 is a d.c. source and may be made any value.

The following table gives sample data on the values of the per cent modulation calculated and measured. This table shows that the accuracy of the measurement as made was not greater than about 4 per cent. The method, however, is sufficiently accurate for many purposes.

It has proved very useful in making quantitative measurements of the percentage modulation in connection with measurements on electron tube properties which involve the degree of modulation. It could be used in a system to indicate continuously the limits of modulation in a radio telephone transmitting station. The circuit could be coupled to the transmitting set and the resistance R_1 adjusted so that with no modulation the plate milliammeter reads zero. When the current is modulated then the milliammeter would show a deflection

TABLE I

E_1	E_2 Eff.	I_r	I_{MOD}	$I_{MOD}-I_r$	Measured Value $\frac{I_{MOD}-I_r}{I_r}$	Calculated Value $\frac{E_2 \times 1.41}{E_1}$
Modulation frequency = 500 cycles						
120	85	14.3	29.1	14.9	1.05	1.00
147	85	17.8	32.2	14.4	0.81	0.82
185	85	23.8	38.1	14.3	0.60	0.64
Modulation frequency = 60 cycles						
135	96	15.5	31.0	15.5	1.00	1.00
135	90	15.9	31.1	15.2	0.96	0.94
	80	15.8	28.8	13.	0.82	0.83
	60	15.9	25.7	9.8	0.62	0.62
	50	15.9	24.	8.1	0.51	0.52
	40	15.9	22.1	6.2	0.39	0.42
154	60	18.2	27.8	9.6	0.53	0.55
142	50	16.4	24.3	7.9	0.48	0.50
126	85	14.3	28.3	14.0	0.98	0.96
Radio frequency = 470 kilocycles						

which could be calibrated in terms of per cent modulation. The modulation then could be kept adjusted so that the milliammeter reads the degree desired. The instrument could be used at any time to measure the exact degree of modulation.

This method gives a fairly accurate measure of the degree of upward modulation. It cannot, however, give any indication of the characteristic of the current when the modulation is downward. It is believed that it is superior to other simple methods of measuring modulation.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

RADIO RECEIVER TESTING EQUIPMENT*

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Summary—*The difficulties of measuring radio receivers and the lack of standardized equipment has made necessary the development and construction of special apparatus for this purpose. Generally speaking, determining the characteristics of the input signal to the receiver, the conditions of receiver operation, and the character of the resulting output will serve as a proper measure of performance. The "transmitter" used in the measuring setup has associated equipment in the way of attenuators, vacuum-tube voltmeters, modulation meter, and monitor and frequency supervision, so as to control rigidly and to know with great accuracy the character of the input signal. The design and usage of these various units are described in detail. Emphasis is placed on the proper design and use of vacuum-tube voltmeters. A novel modulation meter is described, being simple and accurate.*

Power measuring panels, screening, and other precautions are noted for maintaining the receiver under test in proper operating condition. The output power may be measured under different conditions of usage. A "distortometer" of new design is described, enabling the distortion and overload of the output to be directly measured.

Proper operation of this test equipment will measure radio-frequency amplification, audio-frequency amplification, and detector sensitivity. It will show the radio-frequency selectivity, the audio-frequency fidelity, the overall fidelity, and the overall selectivity. The frequency range of the receiver may be measured. The "hum" voltage of the output may be noted. The maximum undistorted output of the receiver and the percentage of overload at any output can be found. The methods of obtaining these data are given in the paper.

I. Introduction

THE performance characteristics of radio broadcast receivers are measured to aid in development, to check the standards of production and to compare competitive receivers. As these performance characteristics are new conceptions in the electrical industry, new definitions must be agreed upon, and new methods of measurement must be devised to evaluate properly these characteristics. Although the radio industry is still moving rapidly forward, ten years of experience have shown the fundamental basis on which radio receiver performance must rest. It was therefore possible for a group of representative engineers in the radio world to meet, compare experiences, and agree upon certain definitions of performance characteristics. The work of the Standardization Committee of the Institute

* Dewey decimal classification: R343. Original manuscript received by the Institute, December 8, 1928.

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of Radio Engineers has almost been brought up to date, and it is expected that their report will shortly be released.

In addition to defining the performance characteristics of a radio receiver, the Standardization Committee has indicated in a general way how these characteristics shall be measured. There are two ways of measuring such characteristics. The first is relative and consists of comparing the output with the input or with some other unit maintained as a standard. The second way is absolute and consists of determining the absolute values of voltages, currents and power involved, and making such comparisons as are desired from the magnitude of the quantities determined. In general, an absolute method of measurement is best, although usually the most difficult.

The Standardization Committee, realizing the dearth of information on such radio-frequency measurements as they have deemed necessary, have not made detailed provision for the actual measuring equipment. The purpose of this paper is to describe in some detail testing equipment capable of measuring the performance characteristics of modern radio receivers, with design conforming to the restrictions of the Standardization Committee.

In addition to the purely technical problems of measuring extreme values of sensitivity, selectivity, etc., each set provides problems of connection, power supply, operation, and subsidiary equipment. The design of measuring equipment had to include such features as would simplify the actual testing of the receiver. Accuracy was, of course, paramount in the design, although simplicity, reliability, freedom from personal equation, and cost were also considered. The construction of this equipment was begun in June and finished in December, 1927, the installation being made in the Receiver Development Laboratories of the Crosley Radio Corporation at Cincinnati. The progress was necessarily slow because of lack of outside information and because each unit had to be tested as built. In June, 1928, the complete equipment was retested and recalibrated, and some changes, which six months' use had indicated were desirable, were made. The equipment has answered every demand made upon it, and is considered by engineers who have seen it to be the very latest in measuring equipment.

II. General

The equipment is divided into two parts, a transmitter and a receiver measuring unit. The transmitter is enclosed in one copper screened room and the receiver equipment in another screened room. Both of these small shielded rooms are enclosed in a third screened

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attenuator, audio-frequency oscillator, audio-frequency voltmeter, audio-frequency attenuator.

The meters are placed in a horizontal panel in front of the units for convenience in adjusting the transmitter. Any unit, or this meter panel, can be removed independently for test or recalibration.

The power supply for the transmitter comes from storage and dry batteries underneath the units. A small 100-watt motor generator outside the booth furnishes plate supply to those units where the

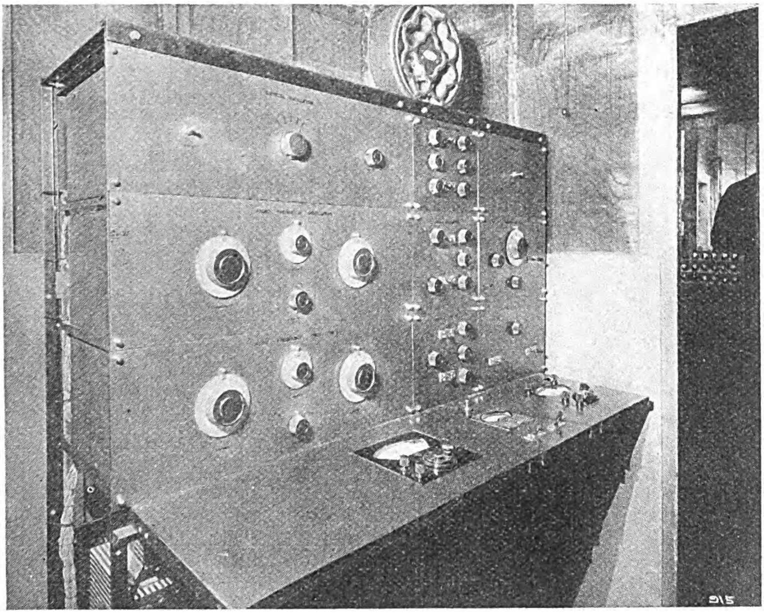

Fig. 2—Transmitter Unit.

plate voltage is not critical. The d.c. output of this generator, properly filtered with respect to commutator ripple and radio-frequency pickup, enters the screen booth through the center of the tube used to connect the screens together and to the ground system. All other external power or control leads, properly filtered, also enter through this pipe, so that the shielding is not impaired.

Fig. 3 shows the receiver testing equipment. The large center panel is for connections and power supply. The units at the ends are a detector calibrating voltmeter, a detector voltmeter, a.c. hum voltage voltmeter, an interference output voltmeter, and a power output voltmeter. The unit at the extreme right is the distortometer, a device

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for measuring the overload and maximum power outputs of a receiver. The unit on the extreme left is a power panel having the proper meters to give 110 and 220 volts at 25 and 60 cycles and 110 and 220 volts direct current.

Batteries to supply the test equipment and to operate battery operated sets are located beneath the bench. Tube racks, tool drawer, desk, and file complete the layout.

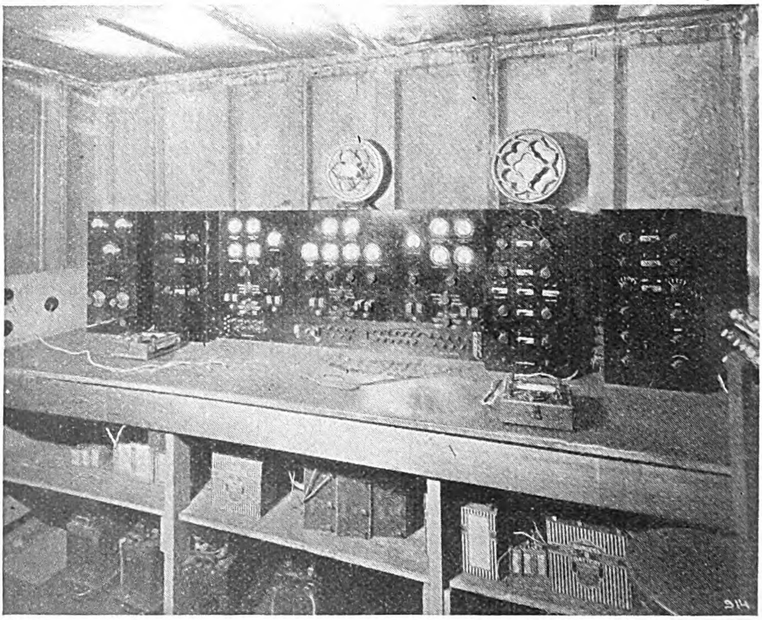

Fig. 3—The Receiver Testing Unit.

III. Radio-Frequency Generation and Measurements

In describing the equipment in detail, the plan will be followed of describing an operation, the reasons for providing a certain unit as constructed, and the means of calibrating and operating. The first and probably the most important operation is that of generating and measuring the radio-frequency voltage applied to the receiver. For some receivers and some measurements this voltage is as low as one millionth of a volt! Considering the fact that this is at radio frequency, it is obvious that straightforward methods of measurement are almost impossible. One method of measurement is to amplify this voltage with a radio-frequency amplifier of known characteristics to a point where direct measurement is possible. This method is often used

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when the source of voltage is not under control, as in field strength measurements.

A second method is a substitution method, but as this essentially means determining the magnitude of the substituting signal, this method is not independent except as applied to field strength measurements. A third method is to obtain this small voltage by the potential drop of a known current through a known impedance. A fourth method is to generate a small voltage by means of a mutual inductance; the frequency, current, and mutual inductance determining the voltage. A fifth method is to measure the voltage at as low a value as convenient and use some form of attenuator of known characteristics to reduce the measured voltage to the desired value. Combinations of two or more of the above methods may be used in some cases.

All of these methods have been tried at various times during the past five years in an endeavor to simplify and increase the accuracy of measurement. The method of measurement obviously influences the methods of power generation, so that the method of voltage measurement must be decided upon first.

The major objections to the first method are that the calibration is not constant as a result of the number of vacuum tubes used and their variations; the calibration varies greatly with frequency; and it is not convenient to change the ratio to conform with widely varying voltages. The principal objection to the third method is its inaccuracy for small voltages. It is impractical to measure currents much below 1 milliampere and inaccurate to use resistances much smaller than 0.1 ohm. This gives a minimum voltage of about 100 microvolts, a figure far too large for very sensitive receivers. Further reduction in the ohmic resistance leads to comparatively large errors due to the difficulty of accurately determining the high-frequency resistance, skin effect, and effective impedance due to reactance. Reactances used as coupling units have the same objections due to resistance. The objection to the fourth method is largely the same as for the third, namely, inaccuracy at low voltages although the large frequency shift is also undesirable. Without resorting to mathematical and form factor calculations (which in themselves are quite susceptible to unknown errors) it is very difficult to obtain a mutual inductance below 5 μh. This gives a minimum voltage of about 1500 microvolts, obviously too high a figure. A major objection to either the third or fourth method lies in the fact that a moment's carelessness on the part of the operator will burn out the measuring thermocouple, and in addition to the cost of a new element, it means delay in replacement

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and recalibration. The principal objection to the fifth method of voltage determination lies in the necessity for providing a sensitive vacuum-tube voltmeter and in the fact that there is a shift in attenuator calibration with frequency unless care is taken in the design.

The choice of a method is therefore determined by the ability to overcome the various difficulties. The design of a stable, sensitive vacuum-tube voltmeter, together with an attenuator independent of frequency, has placed the advantages of the fifth method far in advance of all others. In reliability, foolproofness, and accuracy this method seems the best possible at the present time. Voltages applied to a receiver can be varied between about 1 volt and one half of one millionth of a volt with a known accuracy of better than 3 per cent.

The Radio-Frequency Generator

The radio-frequency generator can next be considered. Due to coupling, both conductive and that resulting from radiated fields, it is desirable never to generate much more radio-frequency voltage than required. For reasons of stability and to conform with the method of voltage measurement chosen, it is necessary to generate a far larger amount of power than is normally necessary. This means that the shielding for the generator and the filter for the power leads must be the best possible. The generator was therefore built inside a copper box with a 1/16 in. wall. All the corners, except the lid, are completely soldered. The lid, with about 1/2 in. of overlapping edge, is screwed to the box with twenty-four machine screws, thus insuring a tight contact. Magnetic flux will leak through cracks that water will not flow through, and as this shielding almost determines the lower limit of voltage measurement, no precaution is too great. This copper box was placed inside a second brass box, forming the unit placed in the frame as shown in Fig. 2. The copper box and the brass box were conductively connected at only one point, thus minimizing the tendency for circulating currents to cause coupling. All ground wires and bypass to ground returns in the circuit are connected as near this ground point as possible, to decrease circulating currents. This practice was followed uniformly throughout the construction of the equipment. The object and usefulness of a shielding metal box is greatly impaired when the box is used as a ground return circuit, because of differences in potentials thus created. Such precautions may seem unnecessary, but the difficulties in removing a few unwanted microvolts resulting from circulating current coupling are otherwise almost insurmountable.

The Standardization Committee has recommended that the frequencies of 600, 800, 1000, 1200, and 1400 kc be adopted as standard,

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and performance characteristics be measured at these points. For some measurements, particularly selectivity and frequency range, higher and lower frequencies must be obtained. The oscillator used has a range from about 500 to 1600 kc.

The circuit is shown in Fig. 4. The approximate values used are indicated on the circuit. The Hartley circuit was used for the oscillator as it is quite stable at all frequencies and is not too rich in unwanted harmonics. A small vernier condenser is provided to give small changes in frequency. There is nothing unusual about the oscillator except

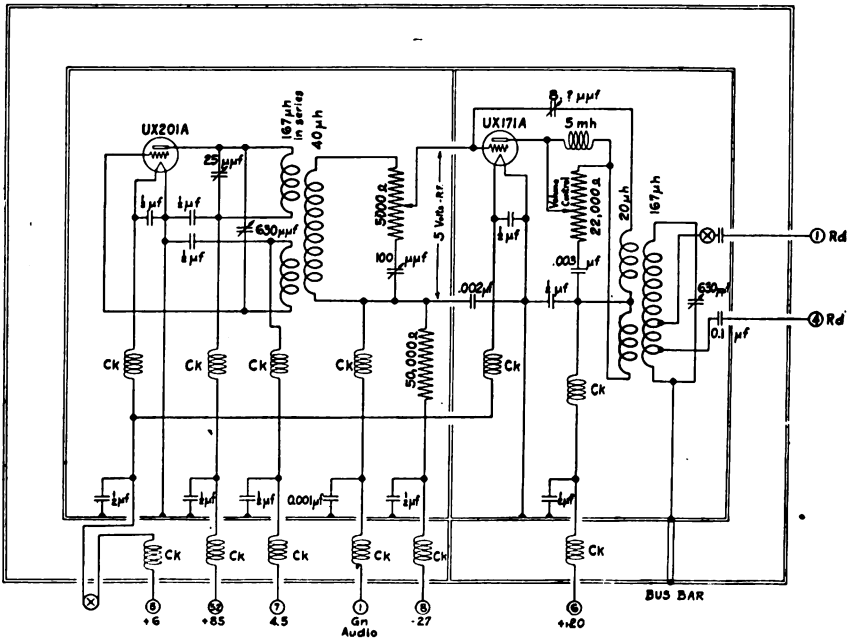

Fig. 4—Radio-Frequency Generator Circuit.

the double filter in each lead but the ground lead. As previously noted, all ground leads and condenser bypass return leads connect to a common ground point, even though the schematic circuit diagrams do not show this. The inductance of each choke is about one millihenry, so designed as to be effective within the range of frequencies used. The condensers used are non-inductive, an important point in using large paper condensers for radio-frequency bypass condensers.

Since the original design in June, 1927, this unit has been made single control, and the oscillator and filter condensers are driven from one shaft. This necessitated some change in the volume control, and that shown in the circuit was finally adopted. At full volume, the

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circuit is a normal amplifier neutrodyned with a "split plate" circuit. As the volume control is moved down, the plate current (r.f.) is "shunted" past the coupling coil. This control works smoothly and gives a wide range of adjustment. The usual volume control methods of filament or plate voltage variation could not be used as this amplifier tube is modulated. It was desired to change the output without changing the percentage modulation, and this could not be done if the operating characteristics of the 171A, used as the modulated amplifier, were changed. It was inconvenient to vary the mutual inductance between the plate coil and the second tuned circuit as this (due to capacity effects) varied the resonant period of the second circuit so much that single control was impracticable. The balance method is somewhat upset by this arrangement, but due to the small coupling back to the oscillator the frequency shift is only about 50 cycles in a million as the volume control is changed. Considering how this control is used such a shift is immaterial.

Grid modulation is used. This is more economical than Heising modulation from the standpoint of available voltage. Tests showed that with a low μ tube, such as the 171A, and with the magnitude of both the radio and audio voltages applied well below the overload point, no distortion due to grid modulation was obtained. In every respect, considering the voltages involved, the grid modulation method is equal or superior to the Heising method as applied in this unit. The variation in percentage modulation is obtained by varying the output of the audio-frequency generator supplying the modulating frequency.

If the magnitudes of the radio-frequency voltage and the audio-frequency voltage are maintained constant on the grid of the amplifier, the percentage of modulation remains constant (within reasonable limits) regardless of the frequencies applied. To keep the radio frequency voltage constant, a resistance of 5000 ohms and a small variable condenser of about 100 $\mu\mu$f were used as shown. The variable condenser was adjusted to give as near a constant value of radio-frequency voltage as possible. This voltage is about 5 volts and varies only about 10 per cent throughout the entire frequency range.

There are two output coils, the values of which are roughly determined by the impedance to which they are supplying power. The lead marked 1*Rd* supplies power to the modulation meter and to the monitor. The lead marked 4*Rd* supplies power to the radio-frequency voltmeter and attenuator and receiver under test.[1] A switch is pro-

[1] The designating numbers used are those on the original layout of the equipment and are here used for convenience in detailing the circuits. All leads with the same numbers are connected. This will aid in tying the various units of equipment together.

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vided inside the inner copper box to open the lead $1Rd$. In the case of extremely sensitive sets, this lead can be opened after the modulation is set, thus reducing unwanted pickup and coupling.

In building this unit, especial attention was paid to loose contacts, vibrating parts, choice of individual tubes used, with the idea of reducing frequency shifting and "wobbulation" with mechanical vibration. The frequency shift is largely determined by the plate voltage, so resistances and voltage measuring posts are provided in the power supply equipment to maintain the oscillator plate voltage absolutely constant. The amplifier plate voltage, not being so critical and requiring more current, is supplied by the motor generator.

The details of calibration and use will be reserved until later under the head of "Frequency Determination."

Vacuum-Tube Voltmeter

Vacuum-tube voltmeters as used throughout in the measuring booth are used so extensively in laboratory measurement equipment that it seems best to include a rather complete description of their theory, use, and misuse. Vacuum-tube voltmeters are roughly divided into three classes. First, the "peak" voltmeter, where the tube and associated equipment are used to indicate the maximum value of the voltage applied. The r.m.s. value, assuming sine wave shape, is obtained by dividing the peak voltage by $\sqrt{2}$. The advantage of the peak voltmeter lies in the fact that the tube used is not calibrated, and in the enormous voltage ranges over which the device will operate. The limits for a high vacuum tube are from about 0.5 volt to 300 volts. When properly used, the peak voltmeter takes no power from the circuit. The disadvantages of this form of voltmeter are its inaccuracy on low voltages, and the large error with bad wave form. This type of voltmeter was used extensively in the development of this equipment and for such work is almost indispensable.

A second type of voltmeter is the simple rectifier type with the tube plate circuit, meter, and source of voltage in series. This has no particular advantage over either the first or third types. It must be calibrated for the particular tube used and forms a load on the circuit voltage of which is to be measured. This load, in the case of high impedance tuned circuits, may destroy the usefulness of the voltmeter.

The third type, ordinarily known as the Moullin type, uses the ordinary three-element tube as a detector, with the grid biased negatively to prevent drawing current and consequently loading the circuit being measured. This type has been extensively described, but an analysis is included here to form the basis for further work on

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the modulation meter later. Also many constructional and circuit changes make it desirable to have the complete development data in one place.

In using the normal three-element tube as a voltmeter several conditions are necessary. The tube is maintained in an operating condition through the proper applied voltages. The grid is maintained negative at all times, grid current never flowing. The entire plate current is due to filament emission, no "gas" effects being present.

The work of Van der Bijl and others has shown that the plate current, I_p, of a vacuum tube is dependent on the grid and plate voltages, and that a grid voltage is μ times as effective as the same plate voltage. μ is approximately a constant and is consequently termed the amplification constant. This relationship can be written as

$$I_p = f\left(\frac{E_p}{\mu} + E_g + \epsilon\right) \qquad (1)$$

where I_p is some function of the plate voltage E_p, the grid voltage E_g and a small voltage ϵ determined by the operating condition of the filament. In general, the interest lies not in I_p but in the change i_p when a small voltage e is applied to the grid. Let the term in parenthesis on the right of (1) be termed E_t. Then

$$I_p = f(E_t). \qquad (2)$$

Now if a small voltage be impressed on the grid of the tube, (2) may be written

$$I_p + i_p = f(E_t + e). \qquad (3)$$

This function may be expanded by Taylor's theorem as follows:

$$I_p + i_p = f(E_t) + \frac{e}{\underline{|1}} f'(E_t) + \frac{e^2}{\underline{|2}} f''(E_t) + \frac{e^3}{\underline{|3}} f'''(E_t) + + + + \frac{e^n}{\underline{|n}} f^n(E_t). \qquad (4)$$

In this equation $\underline{|n} = 1, 2, 3, \cdots n$, and $f', f'', f''', \cdots f^n$ are the first, second, third, etc., differential coefficients of I_p with respect to E_t.* These derivatives may be written in terms of I_p and E_t as follows:

$$f(E_t) = I_p\,;\ f'(E_t) = \frac{dI_p}{dE_t};\quad f''(E_t) = \frac{d^2I_p}{dE_t^2};\quad f'''(E_t) = \frac{d^3I_p}{dE_t^3}\,.$$

* The differential coefficients of I_p should be taken with respect to e, but due to the symmetry of the function the derivative obtained is the same as that with respect to E_t.

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In substituting these values in (4) the initial value I_p may be cancelled and only the change in current i_p retained. Thus

$$i_p = e\frac{dI_p}{dE_t} + \frac{e^2}{2}\frac{d^2I_p}{dE_t^2} + \frac{e^3}{6}\frac{d^3I_p}{dE_t^3} + + \frac{e^n}{\underline{|n}}\frac{d^nI_p}{dE_t^n} + . \qquad (5)$$

This is the fundamental equation used in most vacuum-tube theory. To evaluate this equation, the successive derivatives must be determined under the operating conditions. That is, the value of $\frac{dI_p}{dE_t}$ will not be the same with an impedance in the plate circuit as when this impedance is removed.[2]

When used as a vacuum-tube voltmeter, the voltage applied to the grid of the tube may be written as

$$e = E_m \sin \omega t \qquad (6)$$

where E_m is the peak voltage and $\omega/2\pi$ is the frequency of supply. In the operating range of the tube, there is usually a considerable portion where only the first and second derivatives have a magnitude great enough to be considered. If an operating point be chosen within this range, all terms beyond that containing $\frac{d^2I_p}{dE_t^2}$ may be dropped from (5) without error. Operating at another point will be considered later. With this fact in mind, substituting (6) in (5) gives

$$i_p = E_m \sin \omega t\frac{dI_p}{dE_t} + \frac{E_m^2 \sin^2 \omega t}{2}\frac{d^2I_p}{dE_t^2}. \qquad (7)$$

The average current is the integral of (7) over one complete period divided by the time for that period. Thus

$$i_{p_{ave.}} = \frac{E_m}{2\pi}\frac{dI_p}{dE_t}\int_0^{2\pi} \sin \omega t d(\omega t) + \frac{E_m^2}{4\pi}\frac{d^2I_p}{dE_t^2}\int_0^{2\pi} \sin^2 \omega t d(\omega t). \qquad (8)$$

Performing this integration and substituting limits gives

$$i_{p_{ave.}} = \frac{E_m^2}{4}\frac{d^2I_p}{dE_t^2}. \qquad (9)$$

For a given operating point $\frac{d^2I_p}{dE_t^2}$ is constant and therefore the plate

[2] A complete discussion of this point, and the means taken to determine the operation of the tube in terms of the constants of the tube and circuit may be found in an article by John R. Carson, Proc. I.R.E., 7, 187; April, 1919.

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current change indicated by a d.c. meter is proportional to the square of the applied voltage. Experimental evidence in a well designed voltmeter supports this conclusion.

In order to show the desirability of proper design, especially relating to wave form errors, the equations can be enlarged somewhat. Suppose that the voltage e is no longer a pure sine wave, but the sum of a fundamental frequency $\omega/2\pi$ and some harmonic X of this frequency. Then

$$e = E_1 \sin \omega t + E_2 \sin X\omega t. \tag{10}$$

Suppose an operating point is chosen with the third and fourth derivatives of (5) having values which are not negligible. Then substituting (10) in (5) gives

$$\begin{aligned} i_p = [E_1 \sin \omega t + E_2 \sin X\omega t]\frac{dI_p}{dE_t} &+ \frac{[E_1 \sin \omega t + E_2 \sin X\omega t]^2}{2}\,\frac{d^2I_p}{dE_t^2} \\ &+ \frac{[E_1 \sin \omega t + E_2 \sin X\omega t]^3}{6}\,\frac{d^3I_p}{dE_t^3} \\ &+ \frac{[E_1 \sin \omega t + E_2 \sin X\omega t]^4}{24}\,\frac{d^4I_p}{dE_t^4}. \end{aligned} \tag{11}$$

The average value of (11) is obtained by integrating with respect to ωt over one complete period (which corresponds to X complete periods for the harmonic voltage) and dividing by the time per cycle. The actual integration will not be given here, since the method is outlined in any table of integrals. Expanding, integrating, and substituting the correct limits results in the dropping of the first and third terms, and simplifies into

$$i_p = \frac{E_1^2 + E_2^2}{4}\,\frac{d^2I_p}{dE_t^2} + \frac{(E_1^2 + E_2^2)^2 + 2E_1^2E_2^2}{64}\,\frac{d^4I_p}{dE_t^4}. \tag{12}$$

In calibrating a vacuum-tube voltmeter, a sine wave source presumably is used. Comparing (12) and (9) shows that there is some question regarding what the voltmeter in the second case actually would read. Assuming for a moment the $\frac{d^4I_p}{dE_t^4} = 0$,

$$i_p = \frac{E_1^2 + E_2^2}{4}\,\frac{d^2I_p}{dE_t^2}. \tag{13}$$

As shown from (9) i_p varies as the voltage squared. Therefore the

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equivalent voltage of (13) is $\sqrt{E_1^2+E_2^2}$. Both E_1 and E_2 should be reduced to effective values to express correctly, but are left as peak voltages to avoid confusion later.

The effective value of the sum of two alternating voltages as read by the usual type of a.c. voltmeter is $\sqrt{E_1^2+E_2^2}$. It is therefore obvious from (13) that if $\frac{d^2I_p}{dE_t^2}$ is very large compared with the value of succeeding derivatives, the voltmeter will read true r.m.s. values of voltage regardless of wave form. Inspection of (12) shows that as $\frac{d^4I_p}{dE_t^4}$ increases (and presumably other derivatives also) the average current ($i_{pave.}$) will correspond in calibration to some r.m.s. voltage far different from that actually applied. Correct design is therefore very important, especially where the wave form is not closely sine wave. Voltages have been measured differing by 50 per cent from the true r.m.s. value, when using an incorrectly designed vacuum-tube voltmeter.

Some of these points in design are to use a type of tube and such operating voltages as to provide an operating range well in excess of the voltage to be measured. Preliminary characteristics are greatly to be desired, but in all cases, a check of the voltmeter reading against the true r.m.s. value should be made.

In general, the normal Moullin type of voltmeter is not sufficiently sensitive. The normal range of this type, using a 201A tube is between 0.7 and 3 volts. By using a more sensitive meter, or neglecting wave form errors, this range can be increased somewhat. This difficulty is corrected in the voltmeters used in this measuring equipment by using a d.c. amplifier. The plate battery for the voltmeter tube is connected between its plate and the coupling resistance of the amplifier tube. As the a.c. voltage is applied to the first tube the plate current increases, raising the grid bias on the second tube and lowering its plate current. A meter in the plate circuit of the amplifier tube serves to measure the applied voltage. A schematic circuit is shown in Fig. 5, where the first tube shown is the detector, and the second tube is the d.c. amplifier. This simple circuit is modified somewhat by the controls necessary to obtain the correct initial setting and to control the voltmeter in operation. The sensitivity is determined by the tubes used, battery voltages, coupling resistances, etc. Details are given later in discussing the individual units.

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The usefulness of such a voltmeter is absolutely dependent on its accuracy. This accuracy is determined by the accuracy of calibration and the faithfulness with which this calibration is maintained. It has been found that so long as the operating point does not shift materially from its initial setting, the *change* in plate current with applied voltage is correctly given by the calibration. This means that even though the plate and grid voltages change as much as 10 per cent, the calibration will still be correct, provided that only the *change* in plate current with applied voltage be measured. To accomplish this a reverse current equal to the initial plate current of the second tube is sent through the meter. Consequently this meter reads only the change in plate current. This has an additional advantage in that it enables a very sensitive meter to be used and lowers the minimum voltage measurement limit.

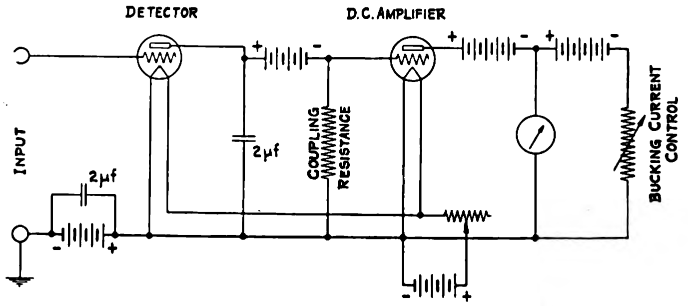

Fig. 5—Vacuum-Tube Voltmeter Circuit.

The biggest variation in calibration is caused by the initial setting, particularly with respect to the filament. Also, as the tube ages its emission characteristics change and destroy the accuracy of calibration. The usual methods of measuring filament voltage or filament current were tried without success. A simple method is to tie the grid and plate together and back to the positive filament terminal through the plate meter. The filament current is then adjusted to give some predetermined emission current, thus setting the tube in the same condition of operation every time it is set. An indication of how reliable this method of adjustment can be is shown in one voltmeter which has been in almost continual use for over three years. Repeated calibrations have shown a steady shift in calibration totaling slightly less than 3 per cent for this entire period. Other voltmeters built since show calibration characteristics indicating even greater constancy.

In determining the emission current initially, the filament current is set about 10 per cent below the normal value and the emission

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current measured. Under conditions of use, a shunt is connected across the indicating meter of such a value as to give full scale reading on the lowest scale (when a multi-scale reading meter is used) when the filament current is properly adjusted. A copper wire wound rheostat is used as filament current control in order to obtain small current changes and consequently accuracy in initial setting.

As the same meter is used to measure the emission and also indicate the voltage, switches marked "start" and "run" are provided to make

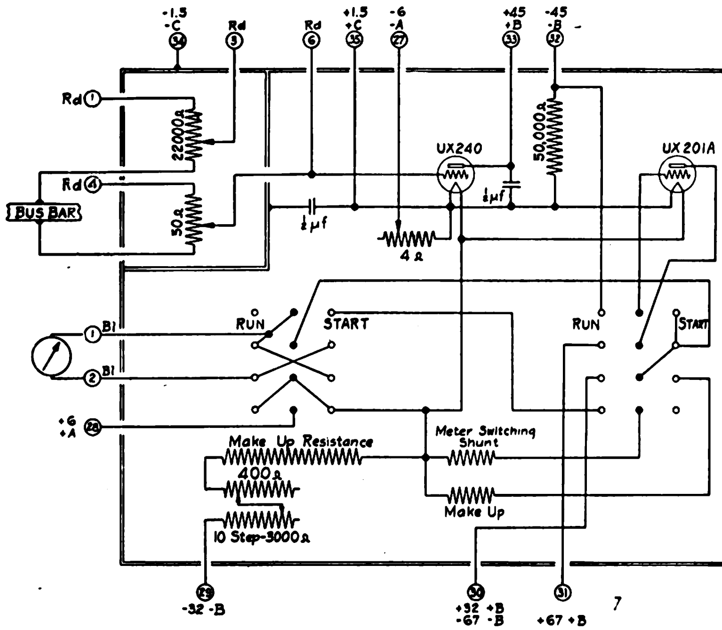

Fig. 6—Radio-Frequency Voltmeter Circuit.

the proper circuit changes. In the case of a two tube voltmeter (rectifier and d.c. amplifier) the second tube only is used to set the emission. To a reasonable approximation, both tubes change in emission characteristics alike and setting one serves to set both.

A voltage about equal to that of the plate battery of the second tube is used for the reverse current battery for several reasons. With a high voltage "bucking" battery, a high series resistance is necessary, and comparatively large changes in resistance only produce small current changes in the reverse current. It is consequently easy to adjust the indicating meter to zero reading, a particularly desirable feature if the meter is very sensitive. Another reason for using similar voltage

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for plate and bucking batteries is that the same type of batteries, having the same current drain, may be used. As a result, the initial zero adjustment of the meter does not have to be greatly changed during the useful life of the batteries. When replacing batteries both units are replaced at once.

The way these various points are taken care of is shown by the circuit diagram and indicated values in Fig. 6. This is the vacuum-tube voltmeter used for the radio-frequency voltages. The sensitivity of this type of voltmeter is determined by the type of tubes used, the value of the coupling resistance, and the sensitivity of the indicating meter. A Rawson multimeter, having four scales of 20.0, 2.0, 0.2, and 0.02 milliamperes at full scale, is used. As indicated, a high-μ tube, UX240, is used as the rectifier with a UX201A as the amplifier. The coupling resistance is approximately 50,000 ohms. To increase the value of $\frac{d^2I_p}{dE_t^2}$, a large bypass condenser is shunted between plate and filament of the first tube, thus increasing the sensitivity. The full scale reading of this voltmeter using the 0.02 scale of the multimeter is approximately 0.1 volt. The minimum voltage readable is approximately 0.02 volt and the maximum (using the 2.0 scale) is approximately 1.0 volt. The advantage of having a 50 to 1 voltage range will be best appreciated by those who have tried to cover large voltage ranges with a thermocouple.

The voltmeter is calibrated on 60 cycles from the 110-volt line, using the proper thermocouples and resistance units. Before calibration of the voltmeter, the thermocouple and resistance units are checked in order to obtain the greatest accuracy possible. All voltmeters used in this booth have a known accuracy of better than ± 1 per cent at any scale reading above 10 per cent of full scale. Proper construction, with repeated checks, have shown these voltmeters to be independent of frequency, being accurate at either radio or audio frequencies. For this to be true, the bypass across the coupling impedance must be effective at the frequencies at which the meter will be used. For calibration purposes, a temporary filter is placed between the rectifier and amplifier to prevent the second tube also from rectifying the applied a.c. voltage and thus giving a false reading. In the case of the audio and 60-cycle voltmeters, this filter is made permanent for the above reason.

A few more notes will be of value to anyone constructing this type of meter. Another stage can be added to increase the sensitivity, but experience has shown that the troubles with drift of the meter, leakage,

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and increasing shift of calibration make this step generally inadvisable. Work with the new screen-grid tubes indicates that a voltmeter built on the same general lines as this type can be made from five to ten times as sensitive if necessary. In the construction of this meter it is extremely important that loose contacts and leakage be avoided. Either of these faults will be characterized by the meter needle jumping very erratically, refusing to stay on zero and otherwise misbehaving. Old "B" batteries, with high and variable resistance, will show the same effect. As a general point in design, too high a plate voltage on either tube will cause "spotty" emission from the filament and the same "jumpy" tendency on the part of the meter needle. Ninety volts is about the usual limit with the 201A type filament, and 67 volts is even more satisfactory. The use of 112 tubes is discouraged; they are not sufficiently stable and constant in characteristics for any form of precise work. A properly operating vacuum-tube voltmeter is a "joy to the eye"; this particular voltmeter will only drift about 2 microamperes in 10 hours.

It usually takes about 5 minutes when first turned on for everything to become settled and for the zero to stay fixed. After this time the voltmeter is in every respect no more trouble than any other type of voltmeter, and is more convenient because of its greater voltage range.

In the circuit of Fig. 5, two potentiometers are shown shielded from the rest of the circuit. The upper one is the supply control to the monitor and the lower one is the supply to the voltmeter and the attenuator. This is the volume control generally used, and in order to cut to zero especial care is taken to ground the lower end. A large bus bar 2 in.×1/4 in. runs along the back of the transmitter, and all units are grounded to this bus bar with copper strips. In the case of the voltmeter, the ground returns of these potentiometers connect to the bus bar at the point of lowest potential. Thus in the "off" position, the "ground" currents (due to the capacity between the transmitter and the screen) flowing through the bus bar do not produce a voltage effective in the input circuit of the voltmeter.

The monitor control was placed in this compartment merely to preserve a neat appearance in the combined unit, and to be handy to the radio-frequency oscillator so that a short high potential lead could be used.

The Radio-Frequency Attenuator

Considering the sensitivity of the voltmeter, it was apparent that an attenuation ratio of 25,000 to 1 was desirable. Assuming a mini-

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mum coupling resistance of 0.1 ohm, this meant a series resistance of 2500 ohms. Difficulties resulting from capacity coupling, skin effect, and changing ratio are almost insurmountable in maintaining the accuracy desired. A multiple stage attenuator was therefore necessary. With two stages a ratio of about 160 was needed, with three stages about 30, and with four stages about 13. For various reasons the three-stage unit was built. As a frequency shift is extremely undesirable, capacitive and inductive reactance must be negligible. A resistance unit, comprising a short length of No. 40 bare advance resistance wire of about 40 ohms resistance, is held between two heavy copper studs. This fine resistance wire has a high value of R/L per

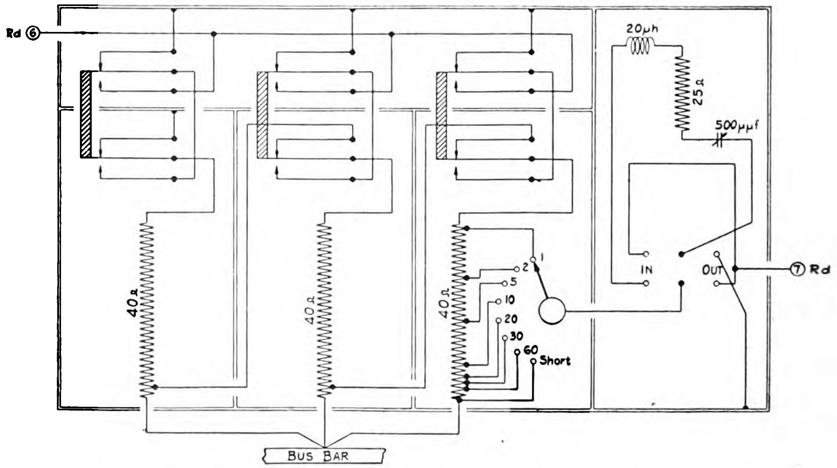

Fig. 7—Radio Attenuator and Dummy Antenna Circuit.

unit of length and a minimum of skin effect. Smaller wire than No. 40 would not have had sufficient mechanical strength. Larger wire would have had a resistance so low that connecting leads would introduce large errors. Heavy copper wedges, with the narrow edge just soldered in contact with the resistance wire, form the taps. Two of the stages have only one tap, at a 30–1 ratio, and the last stage is tapped and connected to a switch giving ratios of 1, 2, 5, 10, 20, 30, 60, and "short."

In a separate compartment in the same unit is included a dummy antenna system. The "standard dummy antenna" is used, consisting of 20 μh, 25 ohms, and 200 $\mu\mu$f connected in series. The condenser is made variable to aid in certain types of development. A switch throws this dummy antenna in or out of the circuit at will.

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The complete circuit is illustrated in Fig. 7 where the stage control switches are schematically indicated. There are five separated compartments in this unit. One holds the dummy antenna, each of three holds an attenuator resistance, and the fifth holds the input lead and switch control shaft. The switches are all normally in the position indicated, and the particular number of stages desired are cut in by operating a cam shaft, depressing the proper switch. The lead which connects the high voltage (comparatively—it may only be 0.1 volt) input to the high side of the attenuating potentiometer is normally disconnected at both ends and grounded. This eliminates any capacity coupling between the input and output. Repeated tests have shown no trace of capacity coupling between input and output or between stages. This is, of course, necessary, or the total attenuation will not be the product of the individual stages, and sufficient range will not be available.

The ground potential ends of the attenuating resistances do not go to the case, but are carried through insulating bushings to a common ground connection on the bus bar. Incidentally, the common ground point is determined by these ground returns; these leads must be short to prevent inductive reactance error.

The attenuator was calibrated at radio frequencies of 500, 1000, and 1500 kc to determine the frequency shift, if any, and to observe capacitive and inductive errors. The proper operating conditions are maintained while calibrating; that is, the last stage is connected across the second, and the second across the first. The common ground point is used as the individual potentiometer ground, and not merely the stud terminal of the resistance wire. The calibration is checked in several ways. The d.c. resistance serves as an approximate check of the ratio, the attenuation at any point being the product of the stages. Another calibration using radio frequency and two vacuum-tube voltmeters gives about the same degree of accuracy. A method which is quite simple and is remarkably consistent and accurate is to take a radio receiver, note the output with a known input voltage and a 1 to 1 attenuator setting. Then change the attenuator to 2 to 1 and increase the input voltage to give the same output. This step by step calibration, with cross checks and reverse calibration, serves to establish the ratios within about 3 per cent error for the worse condition i.e., three stages. The accuracy for the last single stage is much better, being within approximately 1 per cent.

The attenuator will be in error by a considerable amount if a low impedance is connected across the output. If an impedance as low as 100 ohms should be placed across the output (no dummy antenna)

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the possible error would be about 10 per cent. (This on the 2 ratio of the last stage.) Smaller impedances would give greater error. However, the loads are seldom so low in impedance; if they are, a correspondingly lower resistance (higher ratio) may be used and all such error eliminated. The output from the attenuator (RD7) goes through a brass tube to an output post on a small panel in the receiver testing room. This is the upper disk seen on the left wall in Fig. 3.

This combination of radio-frequency generator, voltmeter, attenuator, and dummy antenna forms the input system to a receiver under test. Using a single stage with 1–1 ratio, the maximum applied voltage is of the order of a volt. With all three stages and a 20–1 ratio, the attenuation (based on the present calibration) is 18,200. The minimum voltage is therefore about 0.04/18200 or 2.2 microvolts.

In measuring a receiver, it is assumed that an antenna with an effective height of four meters will be used. The Standardization Committee has chosen such constants for the dummy antenna as will approximate the constants of an antenna with four meters effective height. Therefore the actual voltage applied is divided by 4 to rate the receiver measured. Thus the present equipment is capable of rating any receiver having a sensitivity value above 0.6 microvolt per meter. The accuracy at this extreme reading is approximately ± 5 per cent. As this figure is normally far below the static level, it is thought this limit will be satisfactory for some time to come.

IV. Audio-Frequency Generation and Measurement

The generation and measurement of audio-frequency power and voltage is quite the same problem as that of the radio frequency, with the exception that less complete shielding and better filtering is needed. As more information regarding audio-frequency voltage generation is available, this portion will not be stressed, except to point out certain differences from other methods. The circuit and approximate values are shown in Fig. 8. Fundamentally, the audio frequency is produced by rectification of the heterodyne between two intermediate frequencies. One of these is fixed at about 140 kc,while the other varies between 100 kc and 140 kc, thus giving an audio beat note between 0 and 40 kc. The reason for this large range will be given later. With the fixed oscillator at the high frequency, it is easy to adjust the shape of the condenser plates of the variable frequency circuit to "spread out" the low frequencies, thus keeping about the same percentage accuracy at any dial setting. A small variable condenser, adjustable from the panel, is placed across the "fixed" oscillator. This has a double purpose. When the variable oscillator dial is set to zero, this small con-

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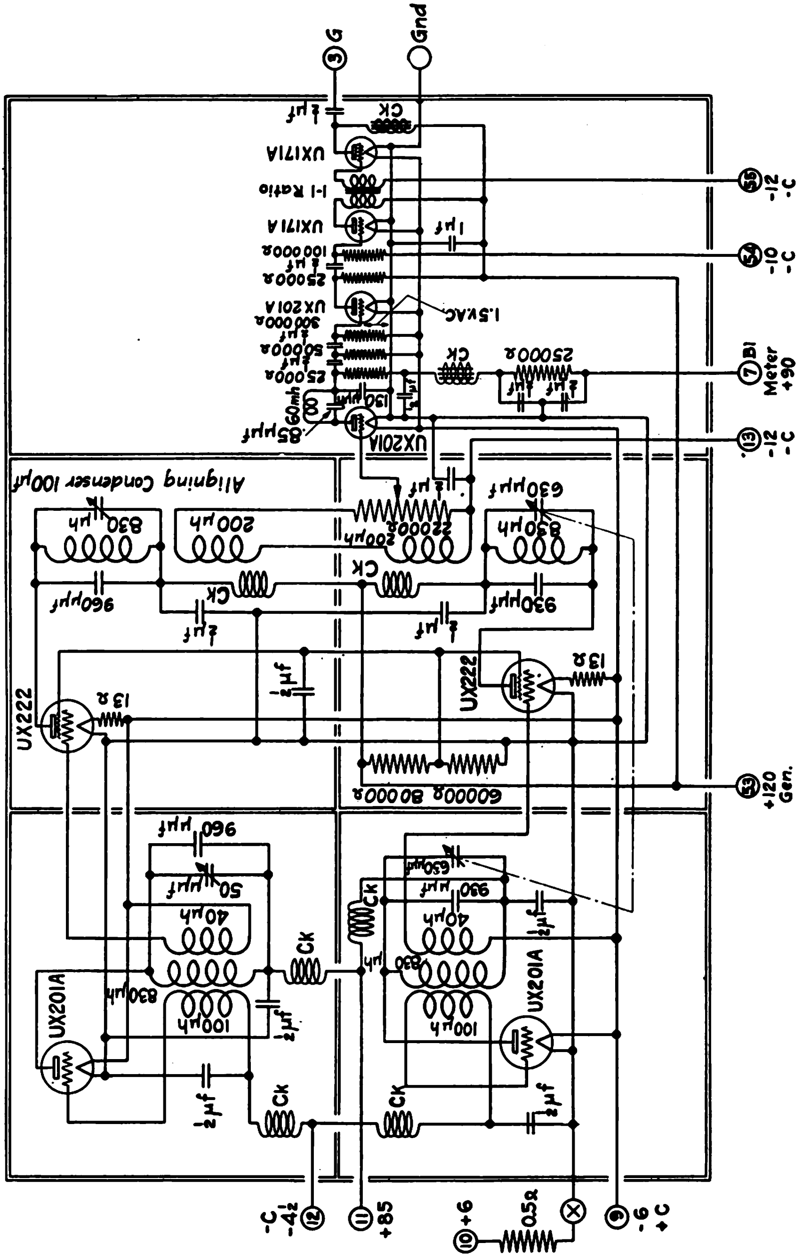

Fig. 8—Audio-Frequency Generator Circuit.

denser may be used to adjust for a true zero beat. This corrects for slight changes in frequency as a result of plate or filament voltages being in error. These small changes are immaterial above a few thousand cycles, but serve to increase greatly the frequency error if not compensated for below this point.

A second use of this small condenser (called a vernier, although it is not such in the usual sense) is to give a wide open scale at the low-frequency end. The vernier dial is set to zero and the variable oscillator dial adjusted to give zero beat. Then, having been calibrated, the small dial serves to give audio frequencies up to about 2000 cycles. The condenser is so shaped that the lower frequencies are well spread out, enabling frequencies of as low as 30 cycles to be set accurately.

As the intermediate-frequency amplifiers are not as selective as the amplifier in the radio oscillator, more care must be taken in regard to harmonics generated in the oscillators. The comparatively smaller frequency variation enables a reverse feedback circuit to be used with stability. The grid coupling coils should be adjusted to give just barely stable oscillation with all condenser adjustments. A slightly greater coupling than this is used to minimize variations resulting from voltage changes.

The oscillators are isolated from the heterodyning circuit by means of the amplifiers. There is no observable tendency to "lock in," stable frequencies as low as one cycle per second being maintained over long intervals. To avoid the necessity of neutrodyning, screen-grid tubes are used as intermediate-frequency amplifiers. Their bias is obtained by a series filament resistance giving about 1.5 volts. The value of the intermediate-frequency voltage is adjusted by means of coupling coils so as to not draw grid current.

The outputs of the amplifiers are coupled through a 22,000-ohm potentiometer, to serve as a volume control. To avoid overloading, the voltage input from each amplifier coil is set about 1 volt. This is about the limit allowable as any greater voltage raises the harmonic content of the output above the permissible value.

Following the detector, a low pass filter circuit designed for a cut off of about 70,000 cycles is placed. This prevents the amplified intermediate frequency from overloading the first audio amplifier. The values given hold only when working between impedances of about 25,000 ohms.

The detector plate current goes through a d.c. meter before returning to the supply source. At low frequencies, say 10 cycles or below, the plate filter is not effective and the meter needle will swing in accordance with the heterodyne frequency. This feature is used to

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set the oscillators to exact zero beat when starting the unit or checking the audio frequency.

In order to pass low frequencies in a resistance coupled amplifier, a high coupling condenser, or a high leak resistance or both, must be used. As it is almost impossible to obtain a large capacity without some leakage, this may place a positive potential on the succeeding grid. This, of course, means grid current and distortion. The leakage problem was solved by using a two-stage resistance capacity coupling, giving a product attenuation for the leakage, yet passing the low frequencies desired without much increase in loss.

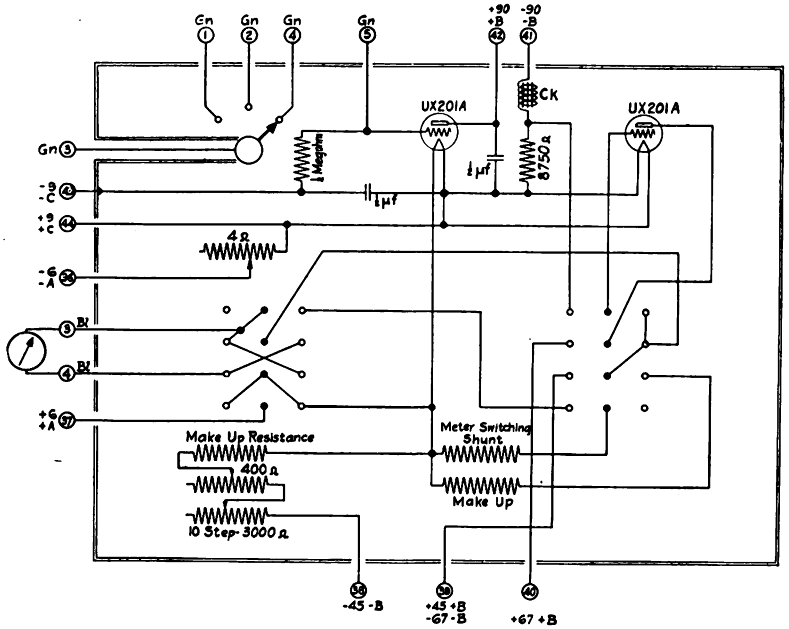

Fig. 9—Audio-Frequency Voltmeter Circuit.

An especially designed transformer is used as the input to the last stage to amplify the low frequencies. A low impedance secondary was used, thus avoiding voltage drop in case of grid current. The grid can therefore "swing" a greater amount, even positive, without causing distortion, and enabling a correspondingly greater power output to be obtained.

The variable oscillator and amplifier are uni-control. The fixed amplifier is broad enough to need no further tuning as the "fixed" oscillator is slightly changed with the vernier. The oscillator tubes are supplied with plate power from batteries with series regulating

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resistances to maintain the voltages constant. The remainder of the unit is supplied from the external d.c. generator.

THE AUDIO-FREQUENCY VOLTMETER

The same method used for measuring the radio-frequency voltage is used for the audio-frequency voltage. The circuit is shown in Fig. 9. It is essentially the same as that of the radio voltmeter, and is included only for the purpose of providing a composite wiring diagram if such should be desired.

The voltmeter needs to be less sensitive and therefore a 201A is used as the first tube and only 8800 ohms for a coupling resistance.

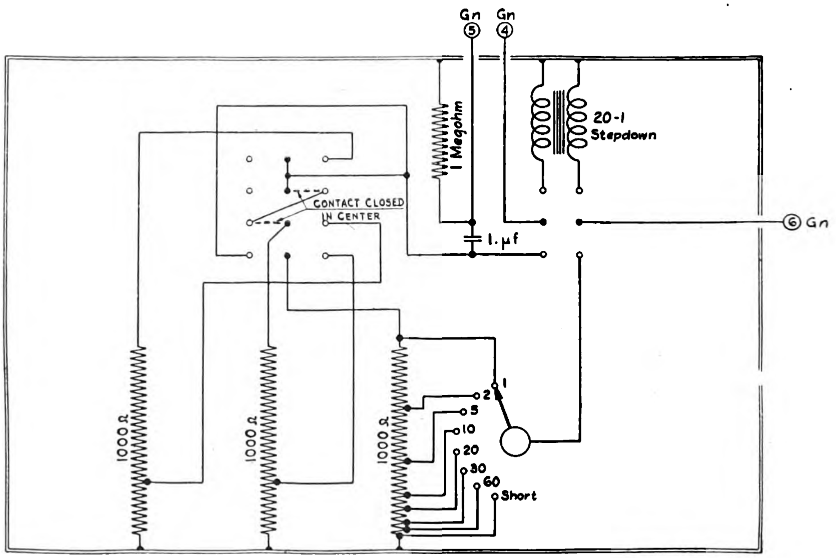

Fig. 10—Audio-Frequency Attenuator Circuit.

The voltage range is from a minimum of about 0.1 volt to a maximum of 2.6 volts. As previously noted, the audio filter is left in the circuit to prevent the audio frequencies acting on the second tube.

The output of the audio-frequency generator goes to a three-point switch located in this unit. These points go to the radio oscillator for modulation purposes, to the monitor, and to the audio attenuator. The only time the audio voltage is measured is when using the attenuator, and therefore only this connection is provided.

THE AUDIO-FREQUENCY ATTENUATOR

The audio-frequency attenuator, shown in Fig. 10, is modeled after the radio-frequency attenuator, having three stages, each tapped

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the same as in the radio-frequency unit. The single stage resistance is 1000 ohms and as capacity effects are relatively unimportant, a Federal anti-capacity switch with properly bent contacts serves as the three-stage switch. A second switch serves to send the audio-frequency generator output through a high ratio step down transformer giving considerable power at a low voltage, as is occasionally desired. The attenuator output lead goes through a brass tube to a contact post similar to that of the radio-frequency attenuator.

V. Frequency Determination

The determination of frequency is a problem second only to that of voltage measurement. An absolute method of frequency determination

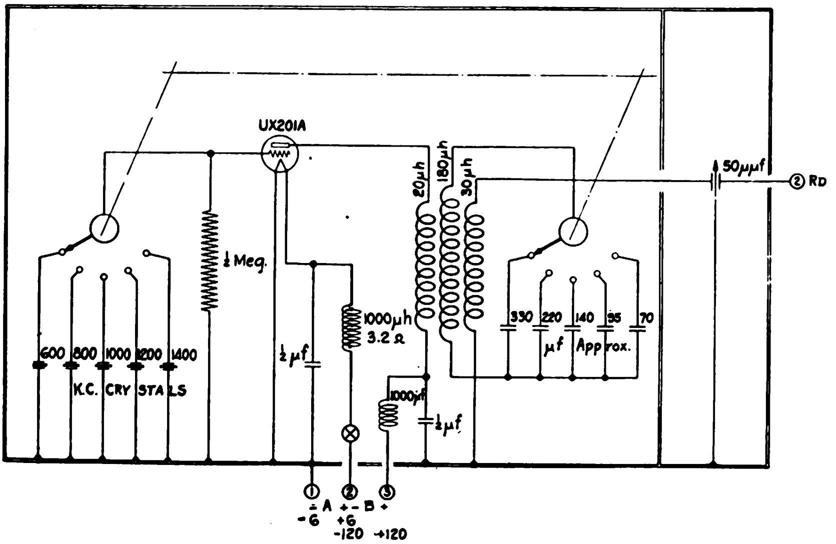

Fig. 11—Crystal Oscillator Circuit.

may be the use of a time and counter checked oscillator with a multivibrator method of frequency multiplication. Such exactness is not required in commercial practice, and therefore no fundamental standard is provided in this measuring equipment. In brief, piezo-electric crystals are used for determining the radio frequencies, and tuning forks are used for determining the audio frequencies.

Two units are provided to aid in this determination. The first is a crystal oscillator shown in Fig. 11. Five crystals with natural frequencies corresponding to 600, 800, 1000, 1200, and 1400 kc are provided. A five throw double-pole switch serves to connect the crystal and the

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corresponding tuned circuit desired. The output is regulated by swinging a grounded rotor between the two plates of a condenser.

There is no temperature control on the crystals. It was not felt necessary to try to maintain the frequency absolutely constant from day to day inasmuch as the frequency shift may only be 1000 cycles under quite wide temperature ranges. The harmonics of the crystals, beat against each other, indicate that the absolute accuracy is not better than ±2 kc. and therefore no further refinement is necessary.

The monitor is the second unit used in determining frequency. The circuit is shown in Fig. 12. It consists of a tube biased to act as a

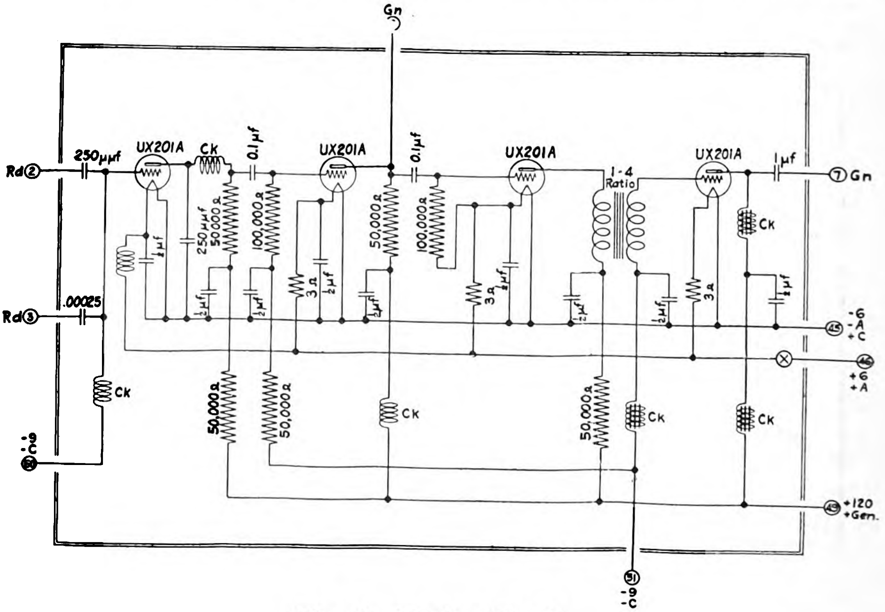

Fig. 12—Monitor Circuit.

detector, a resistance coupled amplifier stage, a second detector, and a second amplifier. The operation is as follows. The output from the crystal oscillator and a portion of the output of the radio-frequency oscillator is impressed upon the first detector of the monitor. By adjusting the radio-frequency oscillator until the resulting beat is zero, the radio frequency applied to the set is the same as that of the crystal. This method is much faster and more accurate than depending upon a calibration curve.

As a result of the detector arrangement, a frequency impressed on the first detector is multiplied and passed on to the second detector where a second beat note may result if another frequency is present.

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Thus 900 kc may be obtained on the radio-frequency oscillator by setting the crystal oscillator to 600 kc and adjusting the radio-frequency oscillator to approximately 900 kc, and then carefully adjusting to make the beat frequency resulting exactly zero. This means that the third harmonic of the crystal oscillator, or 1800 kc, is beating the second harmonic of the radio-frequency oscillator, also 1800 kc, in the second detector of the monitor. This process may sometimes be carried as high as the seventh harmonic, and results in about 30 calibrating points for the radio-frequency oscillator. Any frequency desired and not given by such a possible harmonic ratio may be obtained by reference to the calibration curve. It is only a moment's work to check this calibration if its accuracy is questioned.

The determination of the selectivity curve of a receiver is a job requiring extreme accuracy. In general, a calibration curve cannot be depended upon to give accurately small frequency changes from an initial carrier frequency. Some form of mechanical verniers are often used, but are quite susceptible to error. The method used here has proved quite satisfactory. The crystal oscillator is set at the frequency at which the selectivity curve is desired, say 1000 kc. The radio-frequency oscillator is also adjusted as previously described to exactly 1000 kc. The receiver is then tuned accurately to this signal and not changed further. Suppose it is desired to obtain the receiver response at intervals of 1000 cycles difference. The audio-frequency oscillator is then set at 1000 cycles, and by means of the switch shown in Fig. 9 is connected to the second detector of the monitor. The vernier dial on the radio-frequency oscillator is then changed and a beat note results between the crystal oscillator and the radio-frequency oscillator in the first detector of the monitor. This beat note passes to the second detector and there heterodynes the 1000-cycle audio frequency, and a second beat note results. By proper adjustment of the radio-frequency oscillator this second beat note may be made zero. When this is done, the radio frequency then applied to the receiver differs from the original carrier frequency of 1000 kc by the value of the audio frequency 1000 cycles, giving 1001 kc. The receiver response is then noted on the output meter. This process can be repeated by swinging the radio frequency to the other side of the carrier, again getting zero beat and having a resulting frequency of 999 kc. Setting the audio oscillator at 2000 cycles and repeating gives two more points. Continuing this process gives the complete receiver response curve with respect to frequency, i.e., the "selectivity" curve. The frequency has the accuracy of the audio oscillator frequency setting, which is much greater than that obtained by adjusting the radio-frequency oscillator by calibra-

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tion. The audio-frequency oscillator was made to go to 40,000 cycles so that the selectivity curves could be carried at least that far accurately. For greater frequency differences, the radio-frequency calibration curve can be relied upon. As a result of this accurate method of setting the frequency differences, curves can be repeated almost exactly even though they are quite sharp.

The audio-frequency oscillator is calibrated at 128, 256, 512, 1024, and 2048 cycles by means of tuning forks. A 60-cycle point can be obtained from the commercial power supply. Higher and lower frequency calibrating points are obtained by using the crystal oscillator, the radio-frequency oscillator, and the monitor in much the same way that the radio frequencies are determined. Thus the beat between the crystal oscillator and the radio-frequency oscillator may be set at 2048 cycles with the tuning fork. As a result of the harmonic frequencies generated in the monitor, a zero beat will be obtained when the audio-frequency oscillator is set at 2048, 4096, 8192, etc. By pushing the beat between the crystal oscillator and radio-frequency oscillator up to match one of these high audio frequencies, a step by step calibration is thus worked out. Cross checks, with odd harmonics, serve to increase the accuracy as much as may be desired.

Recalibration at frequent intervals is necessary as there is no fixed frequency to check against as in the case of the radio-frequency oscillator. The frequency shift observed in use has been very small, being approximately 500 cycles in 40,000 cycles after six months' use.

VI. Modulation Measurement

A radio transmitter is essentially a radiator of energy so modulated as to conform to the characteristics of the desired frequency of transmission. The amount of this modulation must vary with amplitude and may vary with frequency. In any case, the response from the receiver will vary with this modulation, and in testing a receiver this amount of modulation must be known.

It may be shown that in any conventional modulation scheme, the energy radiated consists of three parts. These parts are transmitted on three separate frequencies. The first is known as the carrier, and is usually the frequency of the radio-frequency oscillator before it was modulated. The second and the third parts are quite similar in that they usually have the same energy content and are transmitted on frequencies symmetrically placed with respect to the carrier and differing from the carrier by the frequency of the modulation. In the mathematical derivation of the formulas connected with this modulation phenomenon, the current or voltage amplitudes of these three

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parts are usually considered rather than the power relations. With this understood, the term "percentage of modulation" can be used without error. Suppose a carrier wave has a reference value of unity, and each other part, termed sidebands, a corresponding value of 1/2. As the relative phases and amplitudes are at all times changing, it is conceivable to have a time when the two sidebands are in phase, that is, add to a value of unity, and are out of phase with the carrier. The net result at such a time is zero. The carrier is then said to be completely or "100 per cent" modulated. There will also be a time when both sidebands and the carrier add in phase, and the net result will be an amplitude twice the value of the carrier.

The percentage of modulation, designated as K, is therefore the ratio of the sum of the sidebands to the amplitude of the carrier. As it is more convenient, both in plotting curves and in the mathematics, peak values of voltage and current are usually used.

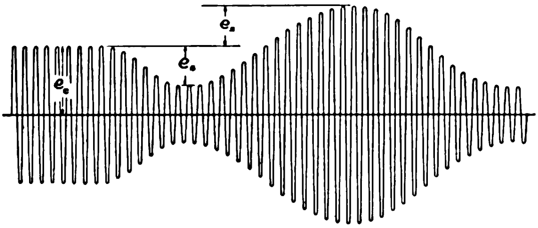

Fig. 13—Modulated Carrier Characteristic.

Fig. 13 depicts how a carrier wave looks before and after modulating. e_c represents the amplitude of the carrier and e_s the sum of the sidebands. By definition $K = e_s/e_c$.

The accurate measurement of K depends on using the curve shape as shown in Fig. 13. The so-called modulation meters on transmitters and broadcast studio control boards are merely meters indicating the relative magnitude of the audio-frequency voltage present. There are several methods of actually measuring K. The first, and by far the most common, is by the use of a cathode ray oscilloscope. An unmodulated carrier wave with an amplitude e_c is impressed on the control electrode. The length of the line on the target represents the amplitude of the carrier wave. The oscillator is then modulated and the new line observed. It will correspond to the maximum value of the envelope or $e_c + e_s$. The difference of the two readings divided by e_c gives K. This method has several disadvantages. The source of the carrier wave should be under control. The cathode ray oscilloscope is expensive. The value of K cannot be determined with much accuracy

because of the necessity for scaling the length of the lines. The biggest disadvantage lies in the fact that the effective value of the carrier e_c may change when modulated. In certain methods of modulation such as that of Heising, the carrier wave may be substantially constant (if properly adjusted), but in other methods the carrier amplitude may change sufficiently to destroy the value of the determination.

If a double element oscillograph is used, the objection based on the change of carrier amplitude may be avoided. The amplitude of the carrier is shown by the distance between two reference lines, one of which is the base line for the superimposed audio frequency. When modulated, the audio frequency is observed as a sine wave rather than merely a trace. The mean of the extreme swings of this frequency is taken as the carrier amplitude, and K is figured as the percentage amplitude of the audio-frequency swing compared with the mean amplitude. If the carrier amplitude does not change when modulated, the mean of the audio swings will remain in the same line as the original base line. The enormous cost of such a unit is prohibitive to many would-be progressive laboratories. And while a visual estimation of K can be made, a scale must be provided and K calculated to obtain any degree of accuracy.

Another method is to use a peak voltmeter to determine e_c and e_c+e_s. This method is subject to practically the same disadvantages as the cathode ray oscilloscope method and is still more inaccurate because of the difficulty of determining the peak voltage accurately.

Another method is to use a peak voltmeter so arranged as to measure first e_c+e_s and then e_c-e_s. The mean of the two readings is e_c effective *while modulated*. This eliminates a large possible error, but is still comparatively inaccurate due to the use of the peak voltmeter. All of these methods have the disadvantage that the percentage of modulation must be calculated. They are in no sense "direct reading."

Considering the use of the measuring equipment, it was evident that none of these methods suited the demands of speed, reliability, accuracy, and direct reading. Based on the theory of the vacuum-tube voltmeter, a design was finally evolved which has proven very satisfactory.

If a radio-frequency carrier of frequency $p/2\pi$ is modulated by an audio frequency of $q/2\pi$ a voltage results expressed by

$$e=E \cos pt(1+K \cos qt) \tag{14}$$

where e is the voltage at a time t, and E is the peak voltage of the carrier wave. If this voltage is impressed on the grid of a vacuum-tube voltmeter with the customary plate and bias conditions, the plate

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current will have a form represented by substituting the value of e given in (14) into (5) as developed for the vacuum-tube voltmeter. This results in

$$i_p = E \cos pt(1+K \cos qt)\frac{dI_p}{dE_t}$$

$$+\frac{E^2 \cos^2 pt(1+K \cos qt)^2}{2} \frac{d^2I_p}{dE_t^2}$$

$$+\frac{E^3 \cos^3 pt(1+K \cos qt)^3}{6} \frac{d^3I_p}{dE_t^3}$$

$$+\frac{E^n \cos^n pt(1+K \cos qt)^n}{\underline{|n}} \frac{d^nI_p}{dE_t^n}. \qquad (15)$$

To obtain all the components of the plate current, (15) must be expanded.[3] For the purposes of this development only the direct current and audio-frequency currents are of interest and all radio-frequency terms will be dropped. The operating point on the tube is so chosen that only the first and second derivatives have any effective value, the remainder of the series being zero. This point is easily obtained in use. With this simplification, the plate current may be shown to be

$$i_p = \frac{E^2}{2} \frac{d^2Ip}{dE_t^2}\left[\frac{1}{2}+K \cos qt+\frac{K^2}{4}+\frac{K^2}{4}\cos 2qt\right]. \qquad (16)$$

The first term inside the bracket of (16) represents the increase in plate current due to the rectified carrier wave. The second term represents the audio frequency of modulation with an amplitude proportional to K. The third term represents an increase in plate current resulting from the sideband input, and is present only when the carrier is modulated. The last term is of double audio frequency, representing the two sidebands beating each other.

If the detector coefficient d^2I_p/dE_t^2 is known for the tube used, and if E and the audio current is measured, K can be calculated. Because of impedance in the plate circuit, the detector coefficient may not be constant with respect to frequency. As the voltage E is modulated, E itself cannot be read accurately. The double frequency term causes error. A more accurate method of determining K may be obtained by breaking (16) into two parts where

[3] The details of this expansion are not given here. The binomial theorem and some trigonometric equations serve to reduce this equation to first order terms of a single frequency.

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$$i_p(\text{d.c.}) = \frac{E^2}{2} \frac{d^2 I_p}{dE_t^2} \left[\frac{1}{2} + \frac{K^2}{4} \right] \tag{17}$$

and

$$i_p(\text{a.c.}) = \frac{E^2}{2} \frac{d^2 I_p}{dE_t^2} \left[K \cos qt + \frac{K^2}{4} \cos 2qt \right]. \tag{18}$$

In the operation of the modulation meter, the modulated input is adjusted to give a constant d.c. change in the plate current, regardless the value of K. Let this value of current be C. Then

$$C = \frac{E^2}{2} \frac{d^2 I_p}{dE_t^2} \left[\frac{1}{2} + \frac{K^2}{4} \right]. \tag{19}$$

From which

$$\frac{E^2}{2} \frac{d^2 I_p}{dE_t^2} = \frac{4C}{2+K^2}. \tag{20}$$

Substituting (20) in (18) gives

$$i_p(\text{a.c.}) = \frac{4C}{2+K^2} \left[K \cos qt + \frac{K^2}{4} \cos 2qt \right]. \tag{21}$$

Inspection of (21) shows that K is absolutely determined if the a.c. component of the plate current is determined. This is obtained by using a vacuum-tube voltmeter and measuring the voltage drop across a known resistance in the plate circuit.

It may be seen from (21) that the current at a frequency $q/2\pi$ is always associated with another current at a frequency $2q/2\pi$, the relation depending upon the value of K. In calibrating the vacuum-tube voltmeter a single frequency is used. Therefore, to read the effective sum of the two terms in the bracket of (21), this must be corrected to give a voltmeter reading independent of the two frequencies present. This may be done by remembering that the effective voltage sum of two voltages is equal to the square root of the sum of their squares. Thus the current is

$$i_p(\text{a.c.}) = \frac{4CK}{2+K^2} \cdot \sqrt{1 + \frac{K^2}{16}}. \tag{22}$$

If K is 100 per cent or 1, the error resulting from not making this correction is about 4 per cent. In general, this is unnecessary.

To complete the calibration of the modulation meter, the audio voltmeter is calibrated and a curve drawn. Some percentage of modulation, say 25 per cent, is chosen to give full scale reading on the audio voltmeter. Having chosen a value of C, in the present case 100×10^{-6}

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amperes, the corresponding plate current may be found to be 48×10^{-6} amperes, peak. Knowing the sensitivity of the voltmeter, say 2 volts, peak value, full scale, the necessary plate resistance is found to be $2/48 \times 10^{-6}$ or 42,000 ohms. Therefore using this value of coupling resistance, the percentage modulation will be 25 per cent whenever the voltmeter indicating meter reads full scale, and providing the input has been adjusted to make C correct. Other percentages of modulation

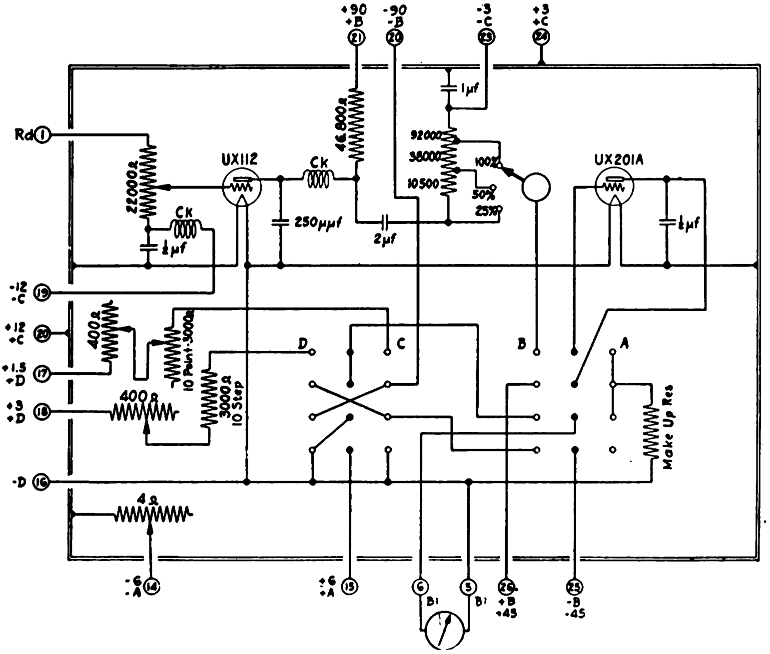

Fig. 14—Modulation Meter Circuit.
Sequence:
A closed, *C* closed, adjust for emission.
B closed, *C* closed, adjust for zero.
B closed, *D* closed, adjust for zero.
B closed, *C* closed, adjust input for *C*.
B closed, *D* closed, read *K*.

can be plotted by assuming K, determining the plate current (a.c.) multiplying by the plate resistance to obtain the audio voltage and then noting the corresponding meter reading.

In use, the plate circuit of the demodulator must be separated from the grid circuit of the audio voltmeter because of the potentials necessary to operate the tubes correctly. Therefore, a resistance-capacity coupling unit is used. The parallel resistance of the combination is used as the effective plate resistance. Taps on the grid resistance

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serve to vary the sensitivity and provide full scale readings of approximately 25 per cent, 50 per cent, and 100 per cent modulation.

It is important to notice in this development that the characteristics of the demodulator do not affect the reading in the least, these characteristics being cancelled out in the simultaneous solution of (17) and (18). A UX201A, a UX171 or a UX240 work equally well as a demodulator, providing sufficient input is available to give the required change C in the plate current. This independence in the determination of K of the characteristics of the demodulator is the surest proof of the accuracy of the method. The cathode ray and peak voltmeter methods are not in themselves sufficiently accurate to check this modulation meter except in a general way. Using the method indicated in the proof, of using d^2I/dE_t^2 as a check, establishes the accuracy of the method.

The circuit and approximate values are shown in Fig. 14. A radio-frequency choke and a small bypass condenser serve to eliminate the radio frequencies before reaching the voltmeter. A single meter, a Westinghouse 100 μa full scale, is used for several purposes. The emission is set on the voltmeter as previously described for accuracy of calibration, and the meter, properly connected and shunted, serves to indicate the correct value. The meter is then connected successively in the plate circuits of the first and second tubes, and the proper "bucking" current controls adjusted to give zero reading on the meter. All calibrations are made from this zero reading position. The meter is then switched back into the plate circuit of the first tube, and the modulated input adjusted with the input potentiometer to give the full scale reading C. The meter is then cut back into the second tube plate circuit where the reading, referred to the calibration curve, gives the percentage of modulation direct.

Through lack of other more accurate means of checking the percentage of modulation, the reading cannot be depended upon much better than within 5 per cent of the value read. The recommended value of K for broadcast receiver testing is set at 30 per cent by the Standardization Committee. The actual value of modulation used is probably between 29 per cent and 31 per cent, a truly negligible error. Recalibration after six months' use showed no change in characteristics, the original calibration checking within less than 1 per cent. Its service has been remarkably consistent and useful.

VII. Radio-Frequency Output Measurement

Having measured the input to the receiver, it is comparatively simple to measure the output. The problem is to measure the output

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without affecting the normal operating characteristics of the set. The only real point of difficulty is in measuring the radio-frequency amplification from the antenna to the detector. Any voltmeter connections across the tuned circuit feeding the detector will change the tuning, increase the feedback to the antenna, and give a false measurement.

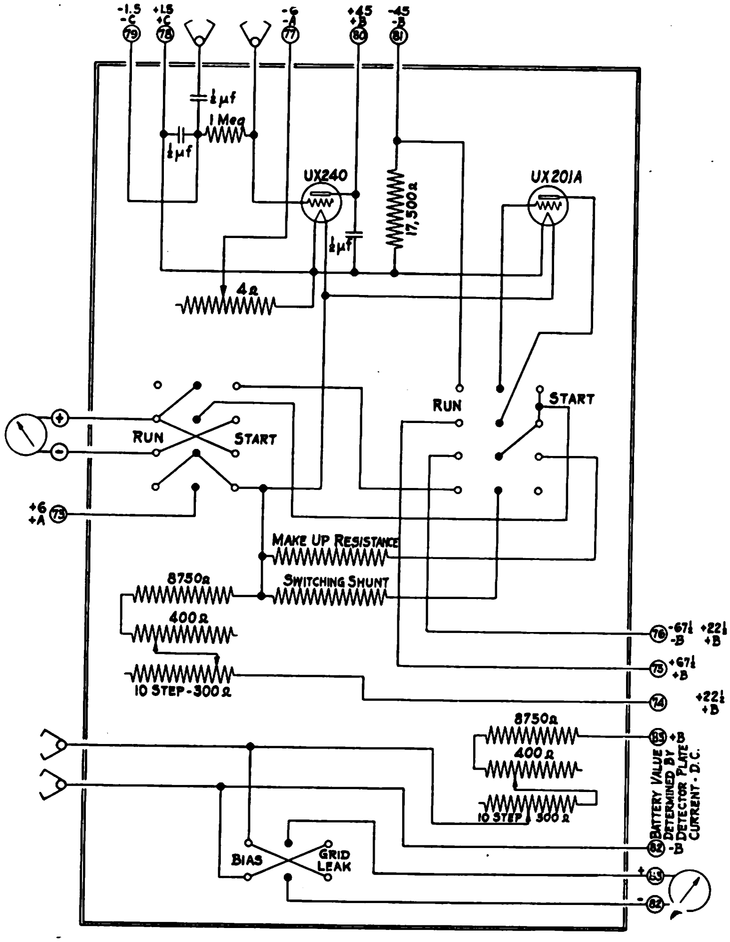

Fig. 15—Detector Calibrating Voltmeter Circuit.

These difficulties can be avoided if the detector tube in the set is used as a voltmeter. It must first be calibrated by noting the change in plate current with applied radio-frequency grid voltage. The current is measured by opening the detector plate lead below any plate impedance which might affect the operation, and inserting a meter. This meter must be fairly sensitive, and a bucking current is provided

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to cancel the initial plate current and bring the meter to zero reading. The radio-frequency grid voltage is measured by temporarily connecting a previously calibrated vacuum-tube voltmeter across the tuned circuit supplying the detector. After the detector is calibrated as a voltmeter, the vacuum-tube voltmeter is removed and radio-frequency

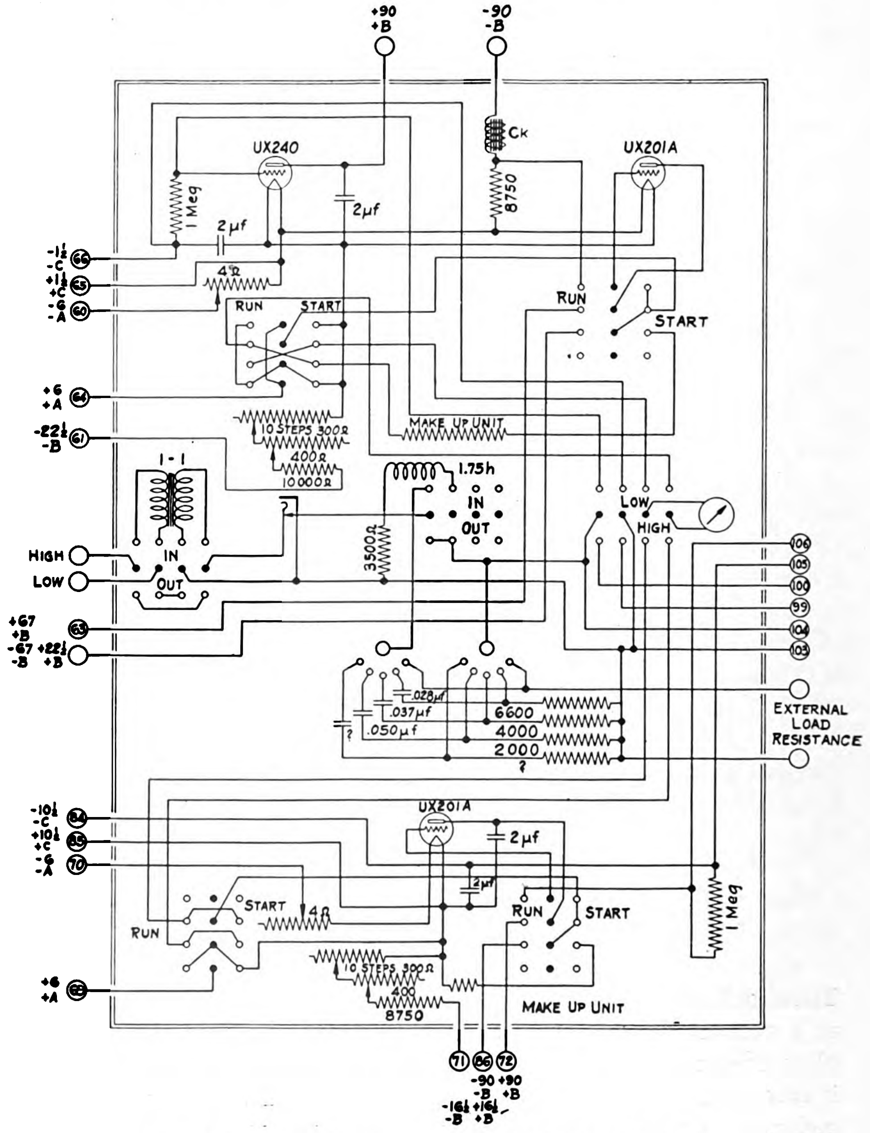

Fig. 16—Audio-Frequency Output Voltmeter Circuit.

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amplification or selectivity can then be measured under operating conditions. Fig. 15 shows the circuit arrangement of the detector calibrating voltmeter and that for the detector-voltmeter. Because of wave form and overload errors with the grid leak and condenser type voltmeter, it is best to operate at low voltages. The detector voltage is usually adjusted to about 1/4 volt for the average set. Connections from the meter panel to the set are made with cords and tip jacks, with spring clips to connect at the proper points on the set.

VIII. Audio-Frequency Output Measurement

Several voltmeters are provided to read the audio-frequency output of the set. One is comparatively low reading, from 0.05 to 1.1 volts, and is used for the measurement of hum and low audio output. The other voltmeter has a range between 0.5 and 200 volts, and is used for interference and normal output voltage measurements. As these are never used simultaneously, a four pole double-throw switch serves to connect the input and output circuits properly so that only one meter is used. Except that the higher reading voltmeter uses only a single UX201A tube, they are similar to the voltmeters previously described. The complete circuit arrangement is shown in Fig. 16.

When measuring a radio receiver, care must be taken to prevent any direct current from the power tube supply system from reaching the voltmeters. If a filter circuit is provided in the receiver, this precaution is unnecessary. If there is no filter supplied, a switch cuts in a 1 to 1 output transformer built to carry the plate current of large power tubes without saturating, and having substantially a flat characteristic from 30 to 10,000 cycles. As recommended by the Standardization Committee, a load impedance equal to the internal plate filament impedance is used. This is connected across the transformer secondary, and the voltage across this load resistance measured. For convenience, five taps are provided giving 2000 ohms for a 171A, 4000 ohms for a push pull 171A, and 6600 ohms for a 120. One tap is left open for future needs to determine the value. The fifth tap connects to external posts on the panel to enable any value of load resistance to be connected.

A measurement of major importance in the development of modern a.c. operated receivers is the hum voltage present in the output. Usually the low reading voltmeter is necessary for this measurement but, unfortunately, the higher reading voltmeter has been necessary on some receivers. Because of the non-linear frequency response of the majority of present day loud speakers, some correction of this hum voltage should be made to represent the audible value of this inter-

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ference. The Standardization Committee recommended a filter system, with values approximately as shown, to make this needed correction. Readings of the hum voltage on a receiver are taken both with and without the hum filter to aid in development or compare receivers. As a different capacity is required with each load resistance, these units are controlled by the same switch that changes the load impedance.

In tuning the set, or adjusting the output, it is often undesirable to have the voltmeters connected. A single closed circuit jack is arranged to connect a loud speaker across the output of the receiver, and opening the voltmeter circuits.

Because of the nature of the output, especially as the overload point is approached, it is important that these voltmeters be carefully adjusted to avoid wave form errors. The values given are approximately right for most tubes, but are checked whenever a tube is changed. The calibration for these meters is accurate to about 1 per cent and is very constant through long periods of time.

IX. Overload Determination

An extremely important measurement in the development and comparison of modern radio receivers is that of distortion. Distortion, or lack of fidelity to the original modulation frequencies, may result from three causes. The output may not contain those frequencies and corresponding amplitudes as were present in the input. The amplitude of the output on any single frequency may not be proportional to the input amplitude. The output may contain frequencies not present in the input.

From the previous discussion it is obvious that proper use of the measuring equipment may quantitatively evaluate the distortion resulting from the first two of these forms of distortion. The measurement of spurious frequencies in the output is somewhat more difficult. Several methods have been used, all having certain advantages and disadvantages. Using a double element cathode ray oscilloscope, and comparing the output voltage with the input voltage, gives one method of measurement. With no other frequencies present, a straight line results on the target. With additional harmonics present, a curved or bent line results. Proper analysis of this line shape gives the individual magnitudes of the various harmonics present. The method is fairly accurate, but rather slow, and requires that an exact replica of the input, in proper amplitude (and preferably in proper phase, or an ellipse results) be available to apply to the oscilloscope.

A second method is to use tuned circuits to pick out the harmonics,

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and measuring their amplitudes individually. This requires sharp tuning circuits, often difficult to obtain at low frequencies. It also means that the entire range of possible harmonics must be measured to obtain the total harmonic power present in the output. The method is rather slow, but has an advantage over the oscilloscope method in that it is not necessary to compare the output with the input.

A third method is to use a frequency bridge to cancel the fundamental frequency and leave only the harmonics. A properly designed "square law" vacuum-tube voltmeter may be used to measure the effective value of all of the harmonic voltages, and therefore evaluate the total harmonic power component. This gives no indication of the relative amplitudes of the individual harmonics, often quite desirable, but serves to measure the total harmonic power output from a receiver.

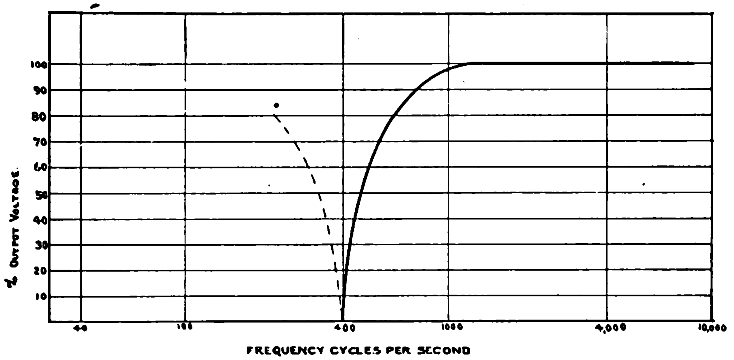

Fig. 17—Distortometer Transmission Characteristic.

The frequency bridge balance method was chosen as being the most desirable in carrying out the purpose of this measuring setup. In order to be effective, and have the accuracy desired, it was necessary that the bridge balance out the fundamental to less than 1 per cent of its original value, and that all frequencies above the second harmonic be transmitted at least 90 per cent. A Campbell frequency bridge circuit was used as the basis of this unit. The ratio between the mutual inductance and the capacity was chosen to give a sharp cutoff near the balance frequency. To accomplish this throughout the audible frequency range, five steps of inductance are used, combined with the proper capacity for balance. The frequency ranges are roughly 30 to 100 cycles, 100 to 300 cycles, 300 to 1000 cycles, 1000 to 3000 cycles, and 3000 to 10,000 cycles.

A typical transmission characteristic is shown in Fig. 17, where the bridge was tuned to 400 cycles. The transmission at 400 cycles is

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negligible, while that at the second harmonic, 800 cycles, is about 92 per cent. Assuming that the greater part of the harmonic power is present at frequencies of the second, third, and fourth harmonics, the average transmission characteristic is about 96 per cent. This assumption will lead to about 4 per cent error in the value measured if all the harmonic energy is concentrated in the second harmonic or in a harmonic above the fifth. The average error resulting from this unknown distribution of harmonic energy will be about 2 per cent of the value read, a truly negligible error.

To make the instrument direct reading in percentage of harmonics, the input is always adjusted to a value of 5 volts, whereupon any harmonic voltmeter reading will always correspond to some specific percentage of harmonic. Thus, if the harmonic voltage component were 5 per cent of that of the fundamental frequency, the vacuum-tube voltmeter would read 0.24 volt. This may be obtained as follows. Assume that the power is concentrated on the fundamental frequency f_1 and any one other harmonic f_n. By assumption (as based on its use) the total applied voltage, 5 volts, equals $\sqrt{E_1^2+E_n^2}$. It is desired to determine E_n when the output contains δ per cent harmonics.[4] Then $E_n=\delta_n E_1$

Solving these two equations gives

$$E_n=5\sqrt{\frac{\delta_n^2}{1+\delta_n^2}}. \tag{23}$$

With 5 per cent harmonic, as assumed above, δ_n is 0.05, and $E_n=0.25$. As a result of the transmission characteristic of the frequency bridge, averaging 96 per cent, the actual voltage reading will be 0.24 volt.

The calculation of sufficient points in this manner and the translation of the actual voltage-meter readings to the distortion factor-voltage calculation enables a direct reading curve to be drawn. The instrument so designed and directly calibrated has been named a "distortometer."

As may be seen from the development of (23), the harmonic voltage E_n is always used as E_n^2. Therefore E_n^2 may be considered as the effective voltage squared, resulting from the summation of the power in all the harmonics, and the method is general in application. That is, the individual harmonics may have any frequency and amplitude relation and the distortometer will still give the correct percentage of effective voltage, referred to the fundamental, in the harmonics.

[4] The symbol δ is here used for the distortion factor and in general use signifies the percentage of voltage, referred to the fundamental, in any harmonic denoted by the subscript. Without the subscript, δ signifies the total effective voltage component of all harmonics in percentage of the fundamental.

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The voltage-meter reading calibration is obtained by impressing known voltages on the input of the distortometer and unbalancing the bridge sufficiently to give practically 100 per cent transmission. The calibration, as may be seen from the circuit diagram, (see Fig. 18) is one which includes the input tube and the bridge circuit as well as the voltmeter tube itself.

The output voltmeter is used in adjusting the input to the distortometer to exactly 5 volts. As a result of this setup, the output attenuator was placed in the distortometer cabinet. The complete circuit of

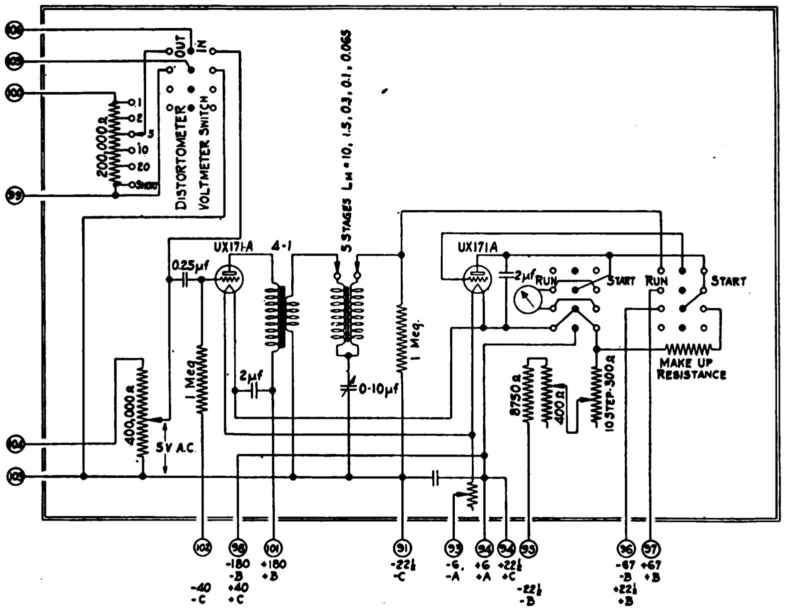

Fig. 18—Distortometer Circuit.

the connections of the output voltmeter and distortometer is shown in Fig. 18. The operation is briefly as follows. The output of the set, at some specified audio frequency, is cut into the distortometer by raising the arm on the 400,000-ohm input potentiometer. The bridge is then tuned by adjusting the mutual inductance and the variable capacity to give a minimum reading in the distortometer output meter. If the output of the set is low, i.e., no harmonics present, this minimum will be practically zero.

The output voltmeter is connected across the input terminals of the unit by throwing the proper switch to the "in" position. The input to the distortometer is then set at 5 volts by proper adjustment of the

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400,000 potentiometer. The reading of the output meter on the distortometer gives the percentage harmonics. If it is desired to determine the power output of the set at some particular value of harmonic, say 5 per cent, the set output may have to be increased. As the set output is increased by proper use of the volume control or by increasing the radio input, the input potentiometer to the distortometer is gradually backed off, keeping the input as indicated by the output voltmeter constant at 5 volts. The output meter on the distortometer will gradually rise, indicating an increasing percentage of harmonic voltage. When the distortometer indicates an output reading of 0.24 volt, the output voltmeter is taken out of the distortometer circuit and cut back across the output load of the set by throwing the switch to the "out" position. This measures the output voltage of the set, and knowing the load resistance, the power output may be calculated. As the distortometer has indicated the percentage of power in the harmonics (0.25 per cent), the set output can be corrected for this amount if desired.

The circuit values are indicated in the drawing and no lengthy comment is necessary. The output voltmeter is adjusted for emission and set as are the other voltmeters described.

To obtain a sharp cut-off on the frequency bridge, it is necessary that the generator supplying the voltage be of comparatively low impedance. Also, this bridge should not load the circuit being measured. A UX171A, having a low plate impedance and followed by a 4 to 1 step down transformer, serves to accomplish both objects. The values of mutual inductance used were determined by the cut-off necessary, their internal resistance, and those coils readily available. The ratio is in all cases 1 to 1. The coils were placed on their cores so that no correction for mutual capacity was necessary.

It is extremely important to use a low flux density in these cores. Saturation otherwise will occur and the harmonics developed in these inductances may be greater than that from the set. This is particularly true on projecting points where crowding of the flux may saturate a small portion of the iron even though the major iron path is quite all right. Until discovered and corrected, this error proved an exasperating problem.

Having a small variable condenser, the capacity range is continuously variable from almost zero to 10 μf, and enables the bridge to be quite accurately set.

If some definite amount of harmonic content is agreed upon as the maximum permissible, it is possible to specify the "maximum undistorted power output" of a receiver. Perhaps a better way of expressing

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Measurement: Form No. [illegible] THE CROSLEY RADIO CORPORATION Date Operators
Receiver Type: Serial No. Engineering Dept. Standard Measuring Booth Data Book No. Page Curve No.

Radio Frequency	Oscillator		Modulation Meter			Voltmeter				Attenuator			Voltmeter			Input Voltage		
	Dial Setting	Frequency	Scale	Reading	% Modulation	Scale	Reading	Voltage		Stage	Tap	Ratio	Scale	Reading	Voltage			

Measurement: Form No. [illegible] THE CROSLEY RADIO CORPORATION Date: Operators:
Receiver Type. Serial No. Engineering Dept. Standard Measuring Booth Data Book No. Page Curve No.

Oscillator		Voltmeter			Attenuator				Voltmeter			Input Voltage	Voltage Amplification			
Dial Setting	Frequency	Scale	Reading	Voltage	Stage	Tap	Ratio		Scale	Reading	Voltage					

Fig. 19—Typical Data Sheets

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this would be the "overload point." Among engineers[5] a tacit agreement of 5 per cent harmonic voltage in the output has been considered as the maximum permissible. Believing the discrimination of the average listener to be somewhat less, the Standardization Committee has suggested that a figure of 10 per cent be used. Because of lack of correlation of aural interpretation with simultaneous measurement of the percentage harmonic present in the output of a receiver, no definite standard has been set. Measurements here indicate that a figure as high as 20 per cent would be even more preferable. In view of the distortion produced in the average loud speaker at various resonant peaks, the setting of a standard of distortion, or permissible distortion factor δ, may well be postponed.

For comparison purposes, measurements at 5 per cent and 15 per cent distortion are being made, thus partially satisfying both viewpoints and also indicating how rapidly the distortion is introduced as the power output is increased.

X. Operation and Conclusions

For convenience in handling all of this equipment, printed data sheets are provided for the recording of the observed data. Two of these charts are reproduced in Fig. 19. Semi-log ruled curve sheets are used for plotting the audio-frequency and fidelity characteristics. Another type of semi-log curve sheets are used for sensitivity and selectivity. All curves are numbered to correspond with the sheet on which the observations were recorded, and with the daily log book describing the purpose and results of the tests. This cross check, together with the index, makes a complete reference system.

In conclusion, a brief summary of the major operations performed by this testing equipment will show its usefulness. By proper manipulation, the outfit will measure radio-frequency amplification, audio-frequency amplification and detector sensitivity. It will show the radio-frequency selectivity, the audio-frequency fidelity, the overall selectivity, and the overall fidelity. The frequency range of the receiver may be measured, and the hum voltage and the effective hum voltage of the output may be determined. The maximum undistorted output of the receiver and the percentage of overload at output can be found. It is a powerful tool in the development of a radio broadcast receiver.

Within the files of the booth described in the preceding pages may be found the performance records of practically every modern radio receiver. The vices and virtues of these receivers are revealed with a completeness that might be shocking to their respective designing

[5] This is largely based on tube distortion calculations.

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engineers. But only as the vices of these receivers are known and corrected, and the virtues seen and appreciated, may the technical operation be improved. It is hoped that this paper may aid in placing the performance rating of a broadcast receiver upon the proper basis, and that from this basis, the improvement in performance may be never ending.

Addendum

In accordance with the idea of keeping this equipment up to date, several minor changes have been made to facilitate the operation since

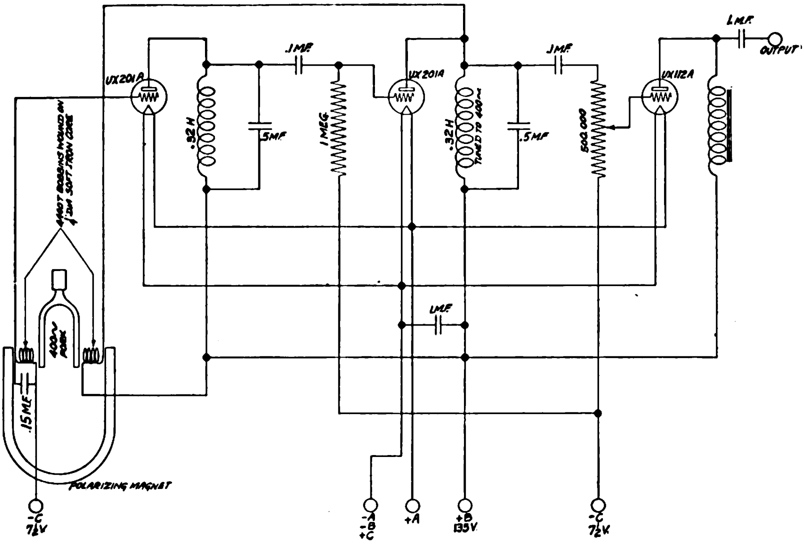

Fig. 20—400-cycle Tuning Fork Oscillator.

this description was written. The advantages demonstrated by the use of a new oscillator are so pronounced that this further addition to the paper seems justified.

Since a considerable portion of the work of measuring the characteristics of a radio receiver is done at 400 cycles, it was felt desirable to install an electrically driven tuning fork to supply power at this frequency. The circuit is shown in Fig. 20, and does not require much comment. The tuning fork used was made by the Waverly Musical Products Corporation. The oscillating circuit consists of the fork circuit, two vacuum tubes, and associated circuits tuned to approximately 400 cycles. Two tubes were found necessary to maintain the proper amplitude and constancy of oscillation. A UX112A as an output tube was found to deliver sufficient power to modulate the radio-

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frequency oscillator to a value of about 60 per cent modulation, and consequently no greater power output system was included. The output contains less than 1 per cent of voltage harmonics, as the wave form is extremely pure. This depends somewhat on the amplitude of oscillation; with the characteristics of the tuned circuits used no amplitude control was found necessary.

The system is rather susceptible to mechanical vibration, and to insure a good wave form the entire unit was mounted on a heavy brass plate and suspended on a soft cushion support. This unit was placed in the same cabinet as the crystal oscillator, some re-arrangement of the parts and controls on the panel being necessary. An output switch (not shown on the diagram) serves to connect the output to the radio-frequency oscillator for modulation or to the monitor for calibration uses, or to the audio attenuator for audio amplification measurements.

The use of this unit, in conjunction with the other apparatus, greatly simplifies the measurement of selectivity. The operation is exactly as previously described, except that the audio oscillator does not have to be re-set on 400 cycles after the radio-frequency has been determined. Extreme accuracy is possible. While valuable perhaps only as a trick, a selectivity curve can be obtained with the radio frequency at the points observed, differing by less than 100 cycles.

Another very valuable use for this oscillator is in the check and calibration of the audio oscillator. As a result of deliberate distortion in the monitor, many harmonics are produced from any applied input. Consequently when the 400-cycle oscillator and the audio oscillator are both connected to the monitor, a great number of "beat" calibrating points are obtained. The detail of calibration is limited almost solely by the ability of the operator to determine the order of the harmonic and thereby the absolute frequency.

It is certain that this oscillator is a valuable addition to the testing equipment, resulting in both a saving of time and an increase in the accuracy of its application.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

MEASUREMENTS OF THE HEIGHT OF THE KENNELLY-HEAVISIDE LAYER*

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Summary—*In this paper we have sought to offer some further contribution to the Kennelly-Heaviside layer problem; first, in the form of experimental data showing clearly evidence of the diurnal cycle in layer height, and secondly, in the form of a discussion of methods for the interpretation of group time and phase retardation experiments and the problem of determining the relationship between the "virtual" and "true" heights. Methods of successive approximation for arriving at the "true" height, from group time or phase retardation measurements, are also discussed and applied. Close accord is found between the results of these methods and the approximation used by Schelleng in a recent paper. The results shown in Fig. 16 also indicate the necessity for further experiments in the important frequency range from 1 to 4 megacycles where no data are available.*

The past few years have contributed much theoretical work and varied experimental data of importance to the problem of determining the properties of the Kennelly-Heaviside Layer and of predicting quantitatively therefrom radio transmission phenomena. It has become evident, however, that the phenomena involved are so many and varied that a considerable mass of experimental data will be necessary before even a semi-quantitative theory of radio transmission will be realized.

Our knowledge of the constitution of the upper atmosphere is so limited that it seems probable that many of its properties will have to be determined by suitably chosen radio transmission experiments before sufficient information will be available for constructing a quantitative transmission theory. Important in this group are experiments dealing with the height of the layer.

One of the objects of this paper is to present a small further contribution to this store of data. However, a brief summary and comparison of accumulated data in the light of recent theoretical contributions, and results readily deducible therefrom, seem appropriate at this time.

As has been frequently pointed out, the phenomena of short-wave transmission are more in the nature of a diffraction than a reflection,

* Dewey decimal classification: R113.4.Original manuscript received by the Institute, December 3, 1928. Presented before meeting of the Philadelphia section, November 23, 1928.

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and the region of the atmosphere of importance in the turning back of waves may be many kilometers in depth. Under these conditions, the terms "layer" and "height of layer" may be misleading unless carefully defined. By "true height of layer" to a ray we will understand the highest point reached by the ray during its trajectory. By "thickness of layer" we will understand the difference between the

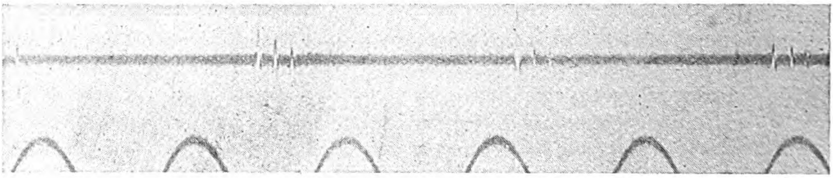

Fig. 1—Transmission from NKF on 4435 kc as Received at Philadelphia on the Evening of September 30, 1928.

height defined above and the lowest height from ground at which the electrical properties of the atmosphere produce an appreciable alteration of the wave trajectory (see Fig. 11). Both of the quantities defined above are, of course, functions of frequency and also the initial angle of the ray.

In one of the major types of experiments which have been used to measure the height of the Kennelly-Heaviside layer, the difference

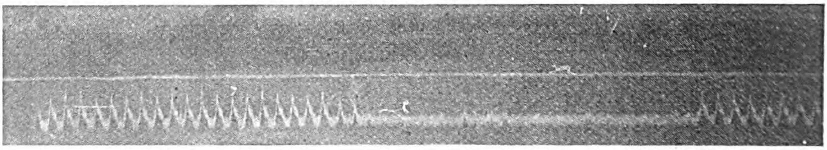

Fig. 2—Transmission with 500-cycle Modulation on October 17, 1928 at 1:45 P.M. Eastern Standard Time.
Note effects of multiple transmission paths.

in group retardation time for a sharp jab or pulse transmitted over two different transmission paths (as, for instance, the ground and first refracted wave paths) is measured.[1] Another of the major types of experiments measures the difference in wave number over several transmission paths.[2]

[1] Breit and Tuve, *Phys. Rev.*, **28**, 554–575; 1926.

[2] E. V. Appleton and M. A. F. Barnett, *Proc. Roy. Soc.* A, **113**, 1926. J. Hollingworth, *Jour.* I. E. E., **64**, 579–595; 1926.

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In each case computations of height have almost universally been arrived at by picturing the transmission to take place between two perfectly conducting, sharply defined planes or spheres at the surface of which the waves were reflected without phase change, although the investigators have in all cases been careful to emphasize that the heights so obtained were not the true heights as defined above. We will term the apparent height thus computed the "virtual height."

It is not a priori evident, for a given layer and frequency, that the results which would be obtained from these two types of experiments

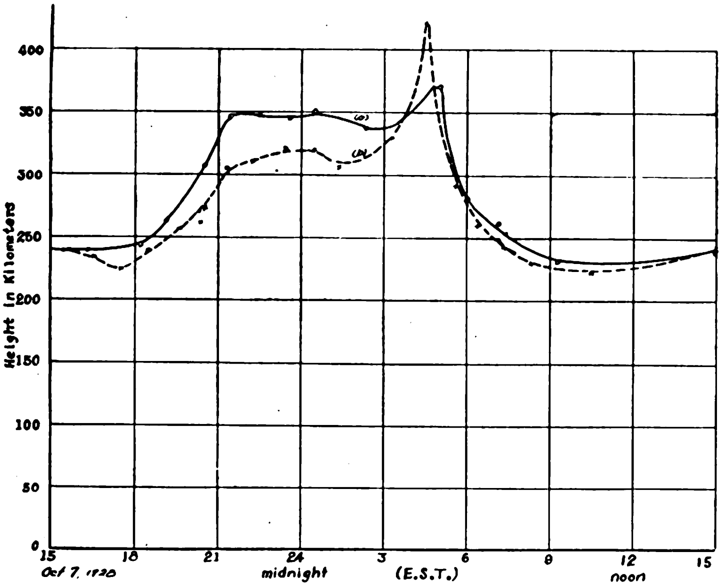

Fig. 3—"Virtual Heights" Obtained during 24-hour Observations, October 7th and 8th, 1928. (a) Observed at Moore School of Electrical Engineering, University of Pennsylvania. (b) Observed at Department of Terrestrial Magnetism, Washington, D. C. Transmission from Naval Research Laboratory, Washington, D. C. on 4435 kc.

would yield the same value for the virtual height, nor is it evident how great a discrepancy exists between the values thus obtained and the true height.

I. Recent Experimental Results

Recent experiments carried on at the Moore School of Electrical Engineering of the University of Pennsylvania in collaboration with the Department of Terrestrial Magnetism and Naval Research Laboratory have furnished further experimental data using the pulse method.

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The transmission was in the form of single sharp, widely separated pulses produced by means of a multivibrator circuit already described in the I.R.E. PROCEEDINGS. The form of the groups of received pulses at Philadelphia (showing clearly the effects of multiple transmission paths) is shown in Fig. 1. In some tests, where 500-cycle modulation was employed, a received signal of the form shown in Fig. 2 was obtained.

In the experiments described here transmission was carried on from the Naval Research Laboratory at Washington (NKF) during alternate half-hour intervals. Simultaneous transmission was furnished from two separate transmitters on the frequencies 4435 kc and 8870 kc. Presumably as a result of skip distance phenomena the higher frequency was not consistently received in Philadelphia

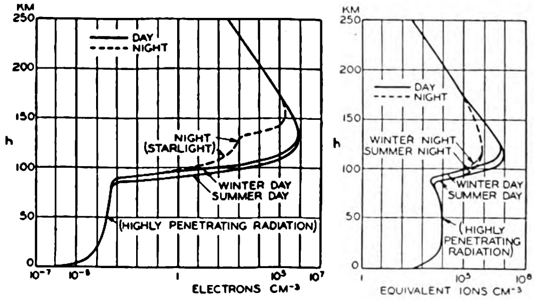

Fig. 4—Electronic and Ionic Densities as Functions of Height and Time (Pedersen).

or at the Department of Terrestrial Magnetism. Faint signals on the frequency 8870 kc were secured when 500-cycle modulation of the form shown in Fig. 2 was employed, but the pulse transmission was not audible or oscillographable at the Philadelphia station.

In Fig. 3 the virtual heights obtained from a 24-hour run on October 7 and 8, 1928, are shown. The full curve shows the results computed from data obtained at Philadelphia, while the dotted curve shows the results reported by the Department of Terrestrial Magnetism. It will be noted that the form of the curves are in essential agreement; i.e., each shows a consistent daytime height rising rapidly during the early evening, then more slowly and finally rising again rather abruptly to a maximum value just before dawn. After sunrise, the curves fall rapidly to the daytime value. A slight tendency for the night values at Washington to lie below those at Philadelphia is indicated by these curves, but the precision of measurement is not at present sufficient to render this difference conclusive for, due to

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oscillographic difficulties, the speed attainable is, at present, limited at Philadelphia to about that shown in Fig. 1. Under these conditions a 10-km difference in height is about at the limit of experimental precision.

II. Methods for the Reduction of Virtual Height Data

With the rapidly increasing store of group time data of the form presented above, and of data involving phase retardation measurements, some further quantitative consideration of the ratio of the virtual heights obtained by means of such measurements to the true height seems appropriate, for such a reduction is an essential pre-

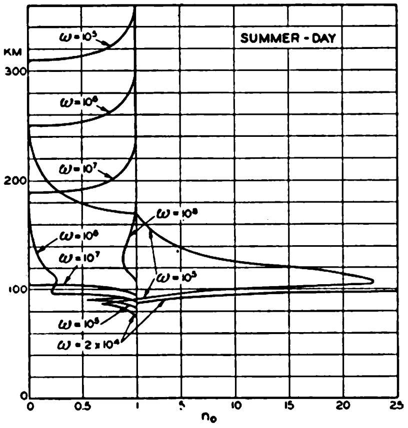

Fig. 5—Refractive Index as Functions of Height and Frequency (Pedersen).

requisite to a determination of the distribution of electronic and ionic densities in the upper atmosphere and its diurnal variation to which the shape of the curves shown in Fig. 3 are supposedly attributable.

Were this distribution known analytically, the problem of determining the ratio of "true" to "virtual" height (which we will denote by h/h') would be analytically formulatable, although the apparent complexity of the problem supposedly renders the chance of the exact integration of the differential equations involved somewhat dubious. In the absence of such data, it is necessary to secure from other sources some tentative evidence as to this distribution upon which preliminary calculations may be based. From these calculations it is possible, given sufficient data from radio experiments, to proceed by methods of successive approximations to an experimental determination of the actual distribution. These methods will be indicated in the course of this discussion.

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After a comprehensive survey of the evidence available, Dr. P. O. Pedersen in a recent book[3] has evolved curves, giving what he considers the best estimates for the variation of electronic and ionic density in the atmosphere as functions of height and time. Some of the curves thus obtained are shown in Fig. 4. While the extremely limited available data upon which these curves are founded necessarily renders them controversial,[4] some of the general properties exhibited are in all probability in accord with the truth.

A substitution of the data on electronic and ionic densities into the well known equations for effective dielectric constant and conductivity as obtained and applied by Nichols and Schelleng, Taylor and Hulburt, Breit and Tuve, and others,[5] gives the index of re-

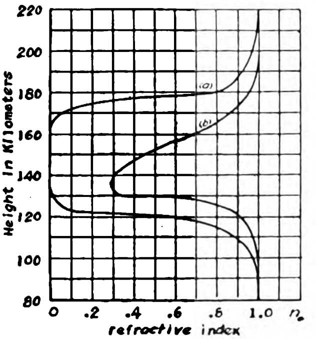

Fig. 6—Refractive Index vs. Height. (a) 4435 kc. (b) 8870 kc.

fraction curves shown in Fig. 5. A curve for the frequencies 4435 kc and 8870 kc is shown in Fig. 6.

As has been frequently pointed out, the elementary considerations of geometrical optics[6] show that, neglecting dissipation and magnetic resonance phenomena, the maximum depth of penetration of a ray does not exceed that at which the refractive index equals the sine of the initial angle of departure of the ray with the vertical. From Fig. 5 it is evident that according to Pedersen's curves for the ionic and electric distribution with height, for $\omega \simeq 3 \times 10^7$, the actual penetration

[3] See "The Propagation of Radio Waves," by P. O. Pedersen. "Danmarks Naturvidenskabelige Samfund," Copenhagen, 1928.

[4] Rather different estimates have been recently arrived at by Hulburt which are not in accord with Pedersen's estimates of temperature or with his assumption of the lack of diffusion in the upper atmosphere. Hulburt's results give much higher values for the height of the layer. See *Phys. Rev.*, 31, 1928.

[5] See Pedersen, page 122. He carefully reviews this material, and extends it to take into account dissipation terms, in many cases omitted in the original treatment.

[6] See p. 175, Pedersen.

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of rays and hence the length of curved path in general is small. It appears possible, therefore, that the curvature of the wave trajectory and errors introduced by the assumption of a triangular trajectory may, in some cases, have received somewhat exaggerated emphasis. A curve showing the height of layer as a function of frequency and ϕ_0 as obtained from Pedersen's n curves is given in Fig. 7.

Let us consider this question in greater detail with a view to estimating the magnitude of departures between real heights and virtual heights estimated on the assumption of a triangular trajectory and

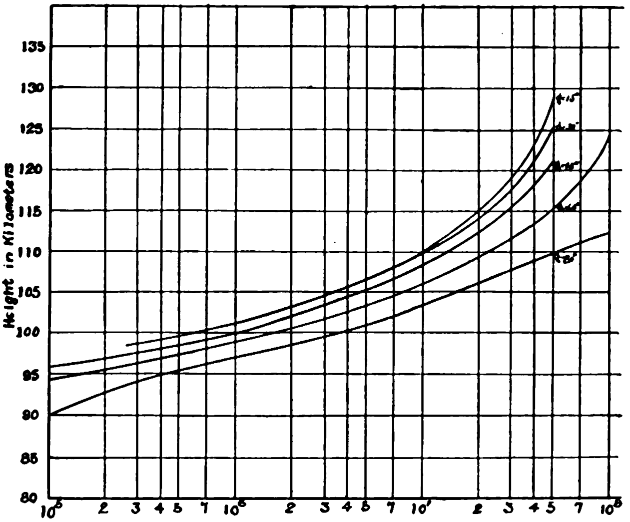

Fig. 7—Height of Layer as Functions of Frequency and ϕ_0 (the initial angle of departure from vertical) as Calculated from Pedersen's n curves (Fig. 5—summer day).

Note that the height thus obtained is probably low.

a velocity of propagation equal to that of light. Breit and Tuve have proved the interesting and important theorem that, in the absence of dissipation or important magnetic effects, the time required for the group signal to traverse the actual curved trajectory is the same as would be required for the group to traverse with the velocity of light a triangular path with the same initial angle of departure as the actual trajectory (see Fig. 8). It is only necessary, then, to compute the equation of the actual path, and to determine the ratio of its maximum ordinate to that of the apex of the triangle in order to compute the value of $\frac{h}{h'}$ for the case of the group time type of experiments.

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III. Differential Equation for the Ray Path[7]

Let us compute, on the diffraction theory, the differential equation of a ray starting at an initial angle ϕ_0 in a medium of gradually varying index of refraction n (which we will subsequently consider as a function of y). We will assume the ray to start at a point where the initial

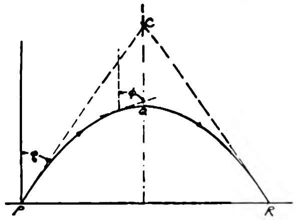

Fig. 8—Equivalent Wave Triangle.

index of refraction is unity. Our fundamental formula of refraction gives us for the angles of incidence and refraction between two media $n_0 \sin \phi_0 = n_1 \sin \phi_1$; when applied to a gradual boundary this becomes[8]

$$n \sin \phi = n_0 \sin \phi_0 = \text{Constant} \qquad \text{(III-1)}$$

Since $n_0 = 1$ in vacuo, (III-1) becomes $n \sin \phi = \sin \phi_0$

$$\frac{n^2}{\sin^2 \phi_0} = \frac{1}{\sin^2 \phi} = 1 + \cot^2 \phi$$

but

$$\frac{dy}{dx} = \cot \phi$$

$$\therefore \quad \frac{dy}{dx} = \sqrt{\frac{n^2}{\sin^2 \phi_0} - 1}. \qquad \text{(III-2)}$$

Which is the required differential equation of our path.

IV. The Variation of n as a Function of y

In order to determine the equation of the ray trajectory in integrated form, we require the analytical equation for n as a function of y. Pedersen's computations of ionic and electronic density from an assumed atmospheric constitution (the results of which are shown in Fig. 4), and the equations which connect these estimates with the refractive index n are sufficiently involved to render small the chance of the determination of a general equation for n as a function of y

[7] See Pedersen, p. 171.

[8] In this analysis we are neglecting earth curvature, an assumption which greatly simplifies the subsequent integrations. Most of the experiments so far carried out are not over sufficient distances to render earth curvature important.

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which will satisfactorily express the entire family of curves for n shown in part in Fig. 5, and at the same time when substituted in equation (III-2) result in integrals which can be evaluated in terms of known functions.

Fortunately, however, the portions of the n curves which are of importance in our calculations of ray paths as a function of y (i.e., the portion corresponding to a decreasing n) are approximated rather satisfactorily by several forms of empirical equations which at the

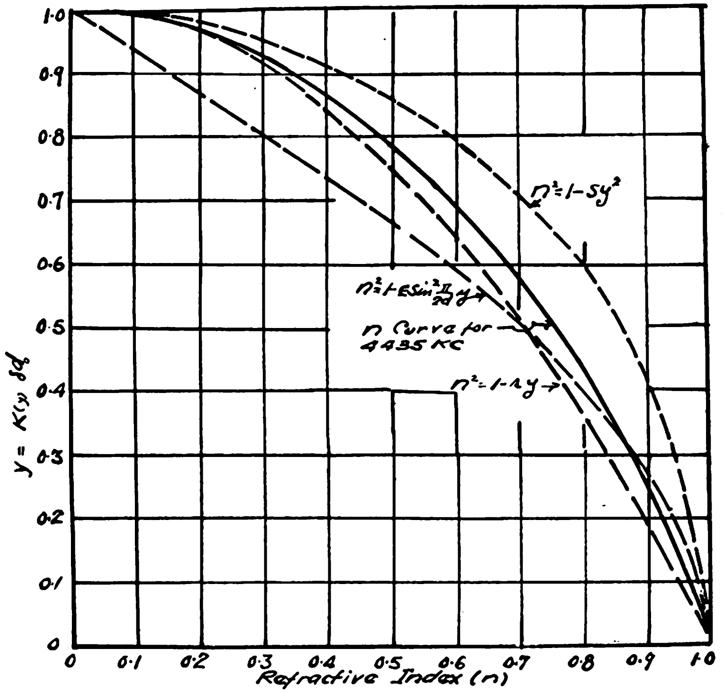

Fig. 9—Showing $K(y)$ (equals $y/\delta d_0$) as a kenetron of n in cases (as (b) and (c) when the constants s, r, E and d, are adjusted to give the same values of $\delta d/_0$ in each case.

same time lead to relatively simple integrals. The three types of equations which we will consider in this paper are

(a) $$n = \sqrt{1 - ry}$$

where r = a constant

y = height above the altitude where n starts to vary appreciably.

(b) $$n = \sqrt{1 - sy^2}$$

where s is another constant and y is the same as in (a).

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(c) $$n=\sqrt{1-E\sin^2\frac{\pi y}{2d}}$$

where E and d are constants.

The form of these empirical curves is shown in Fig. 9.

V. Trajectories Obtained from the Empirical Equations for n

(a) $$n=\sqrt{1-ry}. \qquad \textbf{(V–1a)}$$

Substitution in (III-2) gives

$$\int_0^y \frac{dy}{\sqrt{\cot^2\phi_0-ry/\sin^2\phi_0}}=x.$$

Which yields on integration the parabolic path

$$\frac{2\sin^2\phi_0}{r}\left\{\cot\phi_0-\sqrt{\cot^2\phi_0-\frac{r}{\sin^2\phi_0}y}\right\}=x.$$

This may be written

$$\frac{\cos^2\phi_0}{r}-y=\frac{r}{4\sin^2\phi_0}\left[\frac{\sin^2\phi_0}{r}-x\right]^2$$

but

$$y_{\max}=\frac{\cos^2\phi_0}{r} \text{ or } n_{\min}=\sqrt{1-\cos^2\phi_0}=\sin\phi_0.$$

This checks the condition pointed out in III, i.e., the ray attains a maximum height h for which $n_h=\sin\phi_0=n_{\min}$. The x coordinate corresponding to the maximum value of y is also obtained as

$$x_0=\frac{2\sin^2\phi_0\cot\phi_0}{r}=\frac{2\sin\phi_0\cos\phi_0}{r}=2y_{\max}\tan\phi_0 \qquad \textbf{(V–2a)}$$

or

$$y_{\max}=\frac{x_0}{2\tan\phi_0}. \qquad \textbf{(V–3a)}$$

In other words the maximum height attained by the ray for this type of variation of n with height (which yields a parabolic path) is just one half of the maximum height attained in a triangular ray trajectory having the same angle of departure.

(b) $$n=\sqrt{1-sy^2} \qquad \textbf{(V–1b)}$$

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As in (a), substitution in (III-2) gives

$$\int_0^y \frac{dy}{\sqrt{\frac{1-sy^2}{\sin^2\phi_0}-1}}=x$$

or

$$\frac{\sin\phi_0}{\sqrt{s}}\sin^{-1}\left(\frac{sy}{\cos\phi_0}\right)=x$$

i.e.,

$$y=\frac{\cos\phi_0}{\sqrt{s}}\sin\frac{\sqrt{s}}{\sin\phi_0}x \qquad \text{(V-2b)}$$

The maximum height attained during the trajectory is again

$$y_{\max}=\frac{\cos\phi_0}{\sqrt{s}} \qquad \text{(V-3b)}$$

the substitution of (V-3b) in (V-1b) gives

$$n_{\min}=\sin\phi_0$$

which again checks our fundamental theorem in III.

The ratio of the actual maximum height attained to the height attained by a triangular trajectory of the same initial angle of departure is from (V-3b)

$$\frac{y_{\max}}{y_1}=\frac{\frac{\cos\phi_0}{\sqrt{s}}}{\frac{\pi}{2}\frac{\sin\phi_0}{\sqrt{s}}\cot\phi_0}=\frac{2}{\pi}. \qquad \text{(V-4b)}$$

In other words for this type of trajectory the maximum height attained by the ray is $\frac{2}{\pi}$ times the maximum height of a triangular trajectory having the same angle of departure.

(c) We still have to consider the interesting form

$$n=\sqrt{1-E\sin^2\frac{\pi}{d}y}. \qquad \text{(V-1c)}$$

Proceeding as in cases (a) and (b), we are led upon substitution in equation (III-2) to

$$\int_0^y \frac{dy}{\sqrt{\cot^2\phi_0-\frac{E}{\sin^2\phi_0}\sin^2\frac{\pi y}{2d}}}=x.$$

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Letting $\frac{\pi y}{2d}=y'$ and performing a few obvious simplifications we are led to

$$x=\frac{1}{\cot\phi_0}\int_{y'=0}^{y'=\sqrt{\frac{1-\sin^2\phi_0}{E}}}\frac{dy^1}{\sqrt{1-\frac{E}{\cos^2\phi_0}\sin^2 y^1}}.$$

Letting

$$\frac{E}{\sin^2\phi_0}\sin^2 y^1=\sin^2 Z$$

we have, changing the variable again,

$$x=\frac{(\tan\phi_0)(2d)}{\frac{\pi}{\sqrt{E}}\cos\phi_0}\int_0^{\sin^{-1}\sqrt{\frac{E}{\cos^2\phi_0}}\sin\frac{\pi y}{2d}}\frac{dZ}{\sqrt{1-\frac{\cos\phi_0}{c}\sin^2 Z}} \qquad \text{(V–2c)}$$

Letting the limit of the integral become as large as possible (excluding imaginaries) we have

$$\sin^{-1}\sqrt{\frac{E}{\cos^2\phi_0}}\sin\frac{\pi y}{2d}=1$$

or

$$\sin^2\frac{\pi y_{\max}}{2d}=\frac{\cos^2\phi_0}{E}=\frac{1-\sin^2\phi_0}{E}. \qquad \text{(V–3c)}$$

This gives on substitution in (V-1c)

$$n_{\min}=\sqrt{1-(1-\sin^2\phi_0)}=\sin\phi_0,$$

again our familiar criterion. Our integral then becomes

$$x=\frac{\tan\phi_0\left(\frac{2d}{\pi}\right)}{\sqrt{\frac{1-\sin^2\phi_0}{E}}}\int_0^{\pi/2}\frac{dZ}{\sqrt{1-\frac{(1-\sin^2\phi_0)}{E}\sin^2 Z}}. \qquad \text{(V–4c)}$$

The tabulation of this complete elliptic integral of the first kind $F(\beta, \alpha)$ as a function of $\beta\left(\text{equals }\frac{\pi}{2}\right)$ and $\alpha=\sin^{-1}\frac{\cos\phi_0}{\sqrt{E}}$ is available in tables or graphs.[3] In this case we have, for our ratio of actual

[3] See "Jahnke Einde Funktiontafeln," p. 53.

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maximum height to the maximum height attained in a triangular trajectory of the same initial angle,

$$\frac{y_{max}}{y^1}=\frac{\dfrac{d}{\pi}\sin^{-1}\dfrac{\cos\phi_0}{E}}{\dfrac{d}{\pi}\,\dfrac{\sin\phi_0}{\cos^2\phi_0}\sqrt{E}\,\cot\phi_0 F\left(\dfrac{\pi}{2};\ \dfrac{\cos\phi_0}{\sqrt{E}}\right)}$$

$$\frac{y_{max}}{y^1}=\frac{\cos\phi_0\sin^{-1}\left(\dfrac{\cos\phi_0}{\sqrt{E}}\right)}{\sqrt{E}\,F\left(\dfrac{\pi}{2};\ \dfrac{\cos\phi_0}{\sqrt{E}}\right)} \qquad \text{(V-5c)}$$

Values of the ratio y_{max}/y' for all three cases are shown in Fig. 10 (for case (c) computed for $\phi_0=45°$).

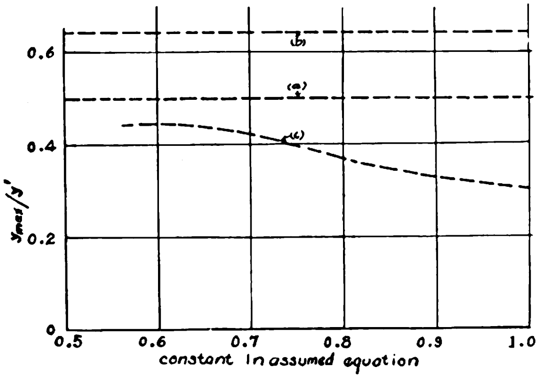

Fig. 10—Showing y_{max}/y'
(a) $n=\sqrt{1-ry}$
(b) $n=\sqrt{1-sy^2}$
(c) $n=\sqrt{1-E\sin\pi/2dy}$

It will be noted that even in case *c* this ratio is surprisingly constant except near the critical angle for which $\sin\phi_0=\sqrt{1-E}$ beyond which rays are not turned back. It will be noted that this form of solution is important in that it gives a minimum value for n and hence exhibits skip distance phenomena which are absent in the other cases since the equation chosen for n goes to zero.

* This is perhaps not the best form of empirical equation for the case chosen in that the minimum given is cuspidal. (for another possibility, see Pedersen p. 160, eq. 24.)

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VI. Computation of the Ratio of True to Virtual Height $\frac{h}{h'}$

The results of the previous section lead us directly to the ratio of true to virtual heights.

It will be noted that we have in each case obtained the ratio $\frac{y_{max}}{y'}$ of the highest point reached in the wave trajectory above the height at which n starts to vary to the distance y' of the virtual triangle (for the group time case) above this same plane.

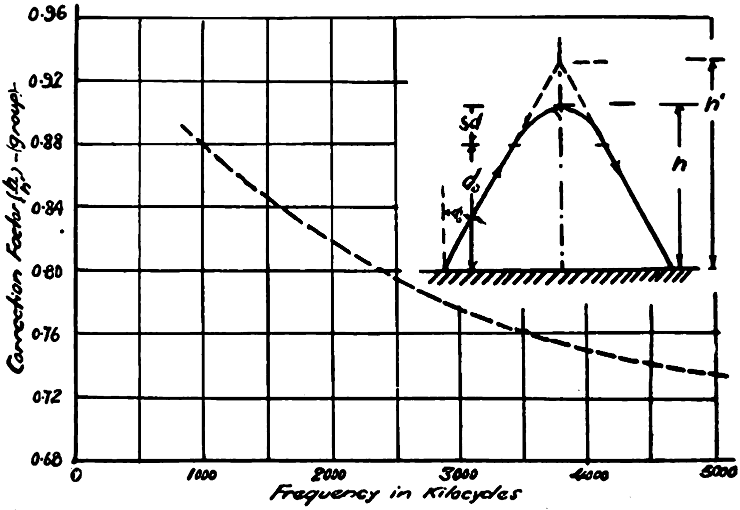

Fig. 11—Correction Factor ($h/h'=1-\delta d/h'$) for "Group Time" Measurements of Heights. Figs. 14 and 15 Are Used for Estimating δd and h', respectively.

As shown in Fig. 12 the total trajectory consists of this path plus two straight curves representing the path of the rays in the lower atmosphere.

Letting $\frac{y_{max}}{y'}=\eta$ it is evident that we have from Fig. 12 by simple geometry

$$\frac{h}{h'}=\frac{d+y_{max}}{d+y'}=\frac{d+y_{max}}{h'} \qquad \text{(VI-1)}$$

or

$$\frac{h}{h'}=\frac{d+y'}{h'}-\frac{y'-y_{max}}{h'}$$

that is

$$\frac{h}{h'}=1-\frac{(1-\eta)y'}{h'} \qquad \text{(VI-2)}$$

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Noting that

$$\eta y' = \delta d \tag{VI–3}$$

this gives us our required result, i.e.,

$$\frac{h}{h'} = 1 - \frac{(1-\eta)}{\eta} \frac{\delta d}{h'} \cdot \tag{VI–4}$$

The factor $\frac{1-\eta}{\eta}$ is readily computable for cases (a), (b), and (c) of Section V by reference to Fig. 10.

It will be noted that the values of $\frac{\eta}{1-\eta}$ in most cases differ but little from unity so that rather closely (exactly in case (a))

$$\frac{h}{h'} = 1 - \frac{\delta d}{h'} \cdot \tag{VI–5}$$

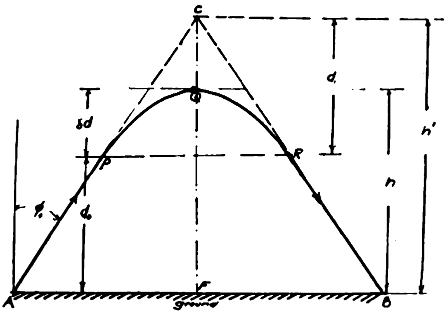

Fig. 12—Showing the Entire Upper Path of Wave Trajectory *PQR* is the Curved Portion of the Path When the Vaying Refractive Index Differs from Unity.

A simple result useful for a rapid visualization of the order of magnitude of the differences between real and virtual heights in any particular case. It will be noted in particular that when δd is small (i.e., when the layer is sharply defined) these heights are in close accord.

The results thus obtained by using the values of δd computed from Fig. 14 are shown in Fig. 11. (See Section VIII.) These curves are necessarily tentative by virtue of the inadequacy of the available experimental data.

As used in this paper, it is evident that the quantity δd is an important one, and careful consideration should be given as to its choice. From this discussion it would appear that δd should be the distance from the lowest point at which the variation in n produces "an appreciable alteration of the wave trajectory." It is evident that a more

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quantitative definition is highly essential. The most satisfactory quantitative definition is to be found in the relation of δd to the constants in the empirical equations used to compute the ray path, thus for equation (V–1a) $\delta d = \dfrac{\cos^2\phi_0}{r}$ etc., In other words the quantity δd (see Fig. 12) should be so chosen that the empirical equation used will give the best possible fit to the n curve. The fit obtained for a particular case is shown in Fig. 9. We denote by δd_0 the value of δd when $\phi_0 = 0$.

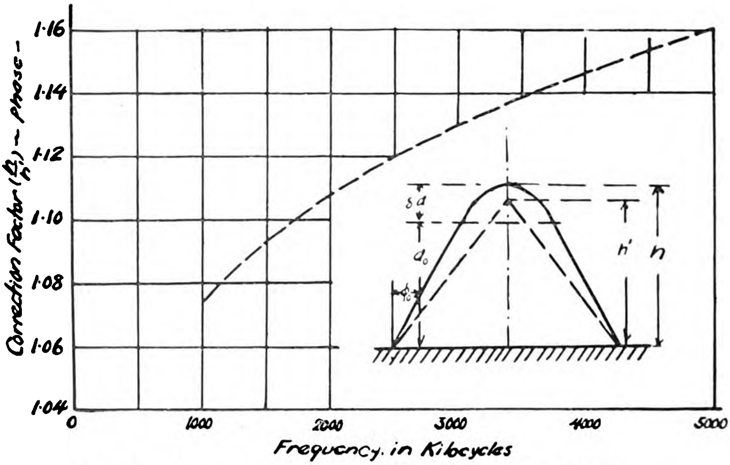

Fig. 13—Correction Factor ($h/h' = 1 - d/3h'$) for "Phase Retardation" Measurements of Heights.

VII. Phase Retardation Experiments

In the previous sections we have shown how the ratio of true to virtual heights of the wave trajectory is computable in the case of group velocity experiments; let us now proceed to a comparison of such results with those obtainable from phase retardation measurements. The procedure in such experiments is to vary the frequency over a small but finite range and to measure

$$\frac{\Delta(N_1 - N_2)}{\Delta f} \qquad \text{(VII–1)}$$

Where N_1 = The number of wavelengths in path a,* let us say the sky path.

N_2 = The number of wavelengths in path b,* let us say the ground path.

* a and b may, of course, be any two transmission paths.

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$$f=\frac{\omega}{2\pi}=\text{transmitted frequency}$$

$$\Delta f=\text{Change of transmitted frequency.}$$

$$\Delta(N_1-N_2)=\text{Change in } N_1-N_2 \text{ due to the change } f.$$

If we define the wave number N by the equation[10, 11]

$$N=fT_p=\int\frac{ds}{\lambda}\cdot \tag{VII–2}$$

Where ds = the element of arc along the path under consideration.
λ = wavelength at the point on the path at which ds is measured.

then

$$N=fT_p \tag{VII–3}$$

and by substitution in (VII-1)

$$\frac{\Delta N}{\Delta f}=\frac{f\Delta T_p+T_p\Delta f+\Delta f\Delta T_p}{\Delta f}\cdot \tag{VII–4}$$

Admitting the approximation

$$\frac{\Delta N}{\Delta f}\backsimeq\frac{dN}{df}\cdot \tag{VII–5}$$

We may show that to this order of approximation the value of $\frac{\Delta N}{\Delta f}\left(\text{equals}\frac{\Delta(fT_p)}{\Delta f}\right)$ approaches the value of T_g, i.e., the group time as measured in group time experiments.

This may be proved as follows:

We have[12]

$$T_g=\int\frac{ds}{u} \tag{VII–6}$$

where: T_g = group time; but also

$$u=\frac{df}{d(1/\lambda)}=\text{group velocity}$$

have

$$T_g=\int\frac{d}{df}\left(\frac{ds}{\lambda}\right)=\frac{d}{df}\int\frac{ds}{\lambda}=\frac{dN}{df}=\frac{d(fT_p)}{df} \tag{VII–7}$$

[10] In the subsequent analysis we will write for brevity N for a certain component (either N_1 or N_2). We will proceed with the analysis for this component. Evidently no generalization is lost and we may proceed to obtain expressions for N_1-N_2 etc. by subtraction of two such results wherever these are pertinent to the discussion.

[11] See Pedersen, p. 175.

[12] Breit and Tuve, *Phys. Rev.*, 28, 571, 1926; equation 10. Also Pedersen, p. 170, equation 46.

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It should be carefully noted, however, that this does *not* imply that T_p is equal to T_g. Actually, if we neglect dissipation it follows[13] that

$$T_p = 1/f \int \frac{ds}{\lambda} = \int \frac{nds}{c} \qquad \text{(VII-8)}$$

while for this case since[14]

$$n^2 = 1 - \frac{4\pi Ne^2}{m\omega^2} \qquad \text{(VII-9)}$$

$$u = cn \qquad \text{(VII-10)}$$

and

$$T_g = \int \frac{ds}{cn} \cdot \qquad \text{(VII-11)}$$

It should be noted that, insofar as equation (VII-5) is justified, phase retardation experiments may be used, by virtue of equation (VII-7), to compute T_g which in turn may be used to compute virtual heights which will, for a given layer, be in accord with those computed from group time measurements. The correction factors deduced in Section VI are evidently applicable to virtual heights thus computed.

In his recent paper Mr. J. C. Schelleng proceeds to use equation (VII-11) in a somewhat different manner: he writes

$$(T_{g_1} - T_{g_2}) = \frac{d(N_1 - N_2)}{df} \qquad \text{(VII-12)}$$

$$(N_1 - N_2)_{f_0} = \frac{1}{f_0} \int_0^{f_0} (T_{g_1} - T_{g_2}) df \qquad \text{(VII-13)}$$

He now identifies the sides of a triangle with the difference of wave numbers $N_1 - N_2$. See Fig. 13.

It should be carefully noted that the altitude of this triangle is in no sense the altitude of the triangle computed from group time data and giving the group time virtual height. This fact is, of course, well appreciated by Mr. Schelleng who presents this triangle as a much closer approximation to the true height of the wave trajectory h. Pedersen has in fact proved that this triangle is indeed a very good approximation; our analysis is in accord with this statement but enables us to find for certain given assumed equations for $n = f(y)$ the exact closeness of accord. Thus we have only to compute the $N = \int \frac{ds}{\lambda}$

[13] See p. 170, Pedersen.
[14] See pp. 174-75, Pedersen.

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over the actual paths given by equations (V-1a), (V-1b), and (V-1c) and then compute the height of a triangle having the same wave number N.

Let us do this for the simplest case; i.e., when $n=\sqrt{1-ry}$ we have the curve represented by Fig. 12, considering the *curved* path of the wave projectile first (i.e., curve PQR (see Fig. 12)) that the number of wavelengths for infinitesimal length ds is

$$dN=\frac{ds}{\lambda}\cdot$$

The number of wavelengths for PQR is hence

$$N=2\int_P^Q \frac{ds}{\lambda}$$

but since

$$n=\frac{\sin\phi_0}{\sin\phi}$$

$$ds=\sqrt{1+\left(\frac{dy}{dx}\right)^2}\,dx=\frac{n}{\sin\phi_0}dx$$

or

$$ds=\frac{n\frac{dy}{dx}dx}{\sin\phi_0\frac{dy}{dx}}$$

assuming

$$n^2=1-ry$$

where $r=$constant

$y=$height (vertical) from P

$$dy=-\frac{d(n)^2}{r}$$

also

$$\frac{dy}{dx}=\sqrt{\frac{n^2}{\sin^2\phi_0}-1}$$

$$\therefore \qquad ds=\frac{-\frac{n}{r}d(n)^2}{\sin\phi_0\sqrt{\frac{n^2}{\sin\phi_0}-1}}=-\frac{nd(n)^2}{r\sqrt{n^2-\sin^2\phi_0}}$$

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Noting that

$$\frac{1}{\lambda}=\frac{n}{\lambda_0}$$

when

$$\frac{\sin\phi}{\sin\phi_0}=\frac{1}{n}=\frac{\lambda}{\lambda_0}$$

$$\frac{N}{2}=\int_P^Q\frac{ds}{\lambda}=-\int\frac{ndn^2}{r\sqrt{n^2-\sin^2\phi_0}}\cdot\frac{1}{\lambda}=-\int\frac{n^2}{r\lambda_0}\frac{dn^2}{\sqrt{n^2-\sin^2\phi_0}}.$$

Let

$$Z=n^2$$

then $Z=1$ at P, and $Z=\sin^2\phi_0$ at Q. Hence

$$\frac{N}{2}=\int_P^Q\frac{dS}{\lambda}=-\frac{1}{r\lambda_0}\int_1^{\sin^2\phi_0}\frac{ZdZ}{\sqrt{Z-\sin^2\phi_0}}$$

$$=-\frac{1}{r\lambda_0}\left[\tfrac{2}{3}(Z-\sin^2\phi_0)^{3/2}+2\sin^2\phi_0\sqrt{Z-\sin^2\phi_0}\right]_1^{\sin^2\phi_0}$$

$$\frac{N}{2}=\frac{2\cos\phi_0}{r\lambda_0}\left[1-\frac{2}{3}\cos^2\phi_0\right]$$

But

$$r=\frac{\cos^2\phi_0}{\delta d}$$

from which it follows that

$$\frac{N}{2}=n\int_P^Q\frac{dS}{\lambda}=\frac{2\delta d}{\cos\phi_0\lambda_0}\left[1-\frac{2}{3}\cos^2\phi_0\right].$$

The above expression gives the number of waves for the curved path; the actual path has, by referring to Fig. 12

$$\frac{N_{\text{total}}}{2}=\frac{(\sec\phi_0)(h-\delta d)}{\lambda_0}+\frac{2\delta d}{\lambda_0}(\sec\phi_0)\left[1-\frac{2}{3}\cos^2\phi_0\right].$$

We wish to find a triangle with the same wave number and same base. The altitude of such a triangle is

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$$h' = \sqrt{(\sec^2 \phi_0)\left[h + \left(1 - \tfrac{2}{3}\cos^2 \phi_0\right)(\delta d)\right]^2 - (\tan^2 \phi_0)(h + \delta d)^2}$$

By solving for the correction factor $\left(\dfrac{h}{h'}\right)$ from the above, we get

$$\frac{h}{h'} = \frac{1}{3}\left(\frac{\delta d}{h'}\right) + \sqrt{1 + \frac{16}{9}(\sin^2 \phi_0)\left(\frac{\delta d}{h'}\right)^2}$$

or by expanding

$$\frac{h}{h'} = 1 + \frac{1}{3}\left(\frac{\delta d}{h'}\right) + \frac{8}{9}(\sin^2 \phi_0)\left(\frac{\delta d}{h'}\right)^2 + \frac{32}{81}(\sin^4 \phi_0)\left(\frac{\delta d}{h'}\right)^4 + \cdots.$$

It will be noted that $\dfrac{\delta d}{h'}$ is a small quantity (usually less than ten per cent) and $\sin \phi_0$ is also small for most experiments. Thus, the product of the squares of two small quantities will be negligible. Hence

$$\frac{h}{h'} \doteq 1 + \frac{1}{3}\left(\frac{\delta d}{h'}\right).$$

This, again, is our familiar expression for $\dfrac{h}{h'}$ (for group case) except for the reversal of sign and the new factor 1/3[15]. In other words the equivalent triangle, which has the same number of waves

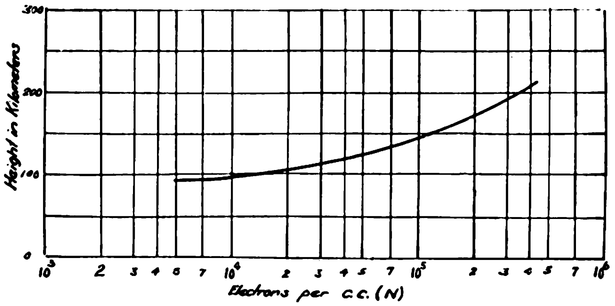

Fig. 14—Electron Density as a Function of Height, as Deduced by the Method of Successive Approximations. Compare with Fig. 4.

as the actual path, has a lower height than the "true" height of the layer. This explains the too large wave number obtained when the triangular height is taken to be the "true" height.[16]

[15] See Section VI.
[16] See Pedersen, p. 176.

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The correction factors obtained from the wave number considerations (using values of δd, estimated from the N and n curves of Figs. 14 and 15) are given in Fig. 13. These factors are evidently highly approximate because of the very limited experimental data on which the N and n curves are based.

VIII. Resumé of Available Data and Method of Successive Approximations

In Sections IV to VI of this paper we have sought to outline methods by which the ratio of "true" to "virtual" heights may be computed in any case in which the equation for the refractive index n is a known function of y; we have investigated the results obtainable

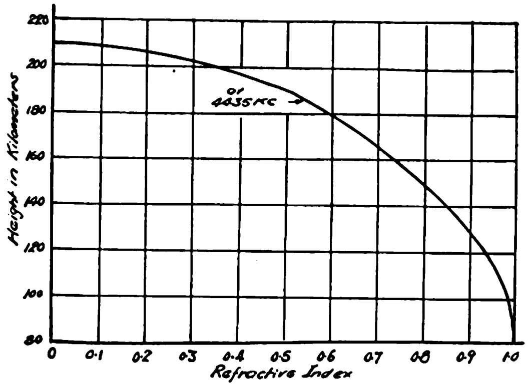

Fig. 15—n as a Function of Height for $f = 4435$ kc.
Note that the curve goes to a higher level than that given in Fig. 6.

by the use of these methods and several convenient empirical equations for n. It is interesting to note that the results of computations on several rather different assumptions give relatively constant and simple relations for the ratio of true to effective height h/h' in terms of the thickness of the n layer δd to which the wave penetrates.

In a recent paper[17] Mr. J. C. Schelleng has presented a brief summary of most of the virtual heights thus far determined by experiment. His estimate of wave number height shows a slow linear variation of this height with frequency. As is emphasized in his paper, the estimate is based on an assumed linearity of the experimental virtual height-frequency curve which is probably not the exact form which will develop when more data becomes available. However, it is interesting to compare value for the corrected height obtained by the

[17] Proc. I. R. E., 16, 1471; November, 1928.

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rather different methods of Sections VI and VII with Mr. Schelleng's estimates. It is interesting to note that the mean height obtained in the experiments described here falls closely on the straight line approximation for the virtual heights as obtained by Schelleng (and used for the virtual height curve in our estimates).

As our first approximation we apply the correction factor $(h/h' = 1 - \frac{\delta d}{h'})$ as obtained from N and n curves derived from this uncorrected (virtual) height curve. From the new height so calculated we can make a second approximation for the electron density N in

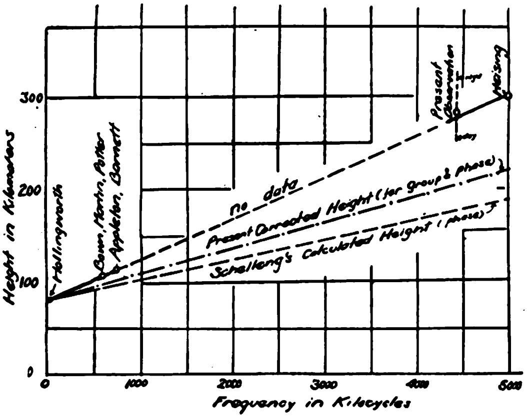

Fig. 16—Showing a Comparison of the Experimental Data, Schelleng's Calculated Height, and the Present Estimate of the "True Height" by Successive Approximations.

the upper atmosphere at different heights. The new N curve enables us to calculate again the curves of index of refraction n for various frequencies and so on. The final curves obtained in this manner (for $f = 4435$ kc) are shown in Figs. 14 and 15. It will be noted that this gives a more gradual variation of n with height, and hence a larger value of the correction factor $\left(1 - \frac{\delta d}{h'}\right)$ than indicated by Pedersen's estimates (see Figs. 4 and 5). The corrected height obtained by this method is shown in Fig. 16. This curve approaches reasonably close to the estimate made by Schelleng; in fact the corrected height obtained from his wave number heights by the application of the appropriate correction factor for this case (see Fig. 13) gives a curve which coincides with the new true height curve to a precision at least as great as warranted by the limited data on which the virtual height curve is based.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

MEASUREMENT OF THE FREQUENCIES OF DISTANT RADIO TRANSMITTING STATIONS*

By
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Summary—*The equipment installed in the Italian Royal Experimental Institute of the Communications for radio-frequency measurements is described. The measurements are based upon the utilization of harmonics generated by the Abraham multivibrator controlled by a 1024-cycle tuning fork. The factors which can influence the variation of the tuning fork, with special regard to the action of temperature, are examined. A simple and rapid method is described, which permits of control of the frequency of the tuning fork at the beginning and at the end of each measurement, and determines, with great accuracy, the frequency of the harmonics between which the wavelength to be measured is included. Such a method consists in comparing the frequency of the tuning fork with that generated by an alternator, and in recording, through a chronograph, the beats included between the two frequencies, the number of revolutions, and frequency of the alternator together with the time furnished by a pendulum of the Astronomical observatory.*

Some examples of calibrations are reported, showing that the degree of precision assured varies from 1 to 5 per 100,000.

The problem of measuring the frequencies of distant radio transmitting stations with the best possible accuracy has recently seriously attracted the attention of workers in the radio field. This problem was not fully appreciated in the past. It was not a rare thing to find wavemeters giving uncertainties of the order of several per cent even in some of the principal radio stations. At the present time the question of precise measurement of the frequencies of radio transmitting stations has become of utmost importance because of the great increase in the number of radio stations, many of which, as for example, broadcasting stations, must be operated simultaneously. As a result, all the most important laboratories are now equipped for radio-frequency measurements of the highest precision.

In Italy the necessity of undertaking special studies and researches in the various government laboratories for the purpose of constructing radio-frequency standards was pointed out last June by Prof. Corbino, President of the Italian Committee of the International Union of Scientific Radiotelegraphy (U.R.S.I.)

Recently the Superior Committee for Vigilance on Italian Broadcasting, presided over by Senator Tittoni, recognizing the necessity

* Dewey decimal classification: R210. Original manuscript received by the Institute, October 15, 1928. Revised manuscript received, November 14, 1928.

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of checking the frequencies of the emitted waves of the Italian broadcasting stations, charged the Ministry of Communications with the execution of these high precision measurements.

The object of this paper is to describe the apparatus used in the Royal Experimental Institute of Communications for the measurement of long and medium waves between 100 and 30,000 meters (3,000 and 10 kc). The system is based on the well known principle of harmonics produced by means of the Abraham multivibrator. The set employed can also be utilized for the measurement of shorter waves, but we believe that measurements of this kind should preferably be made with other methods, the description of which will be contained in a later paper.

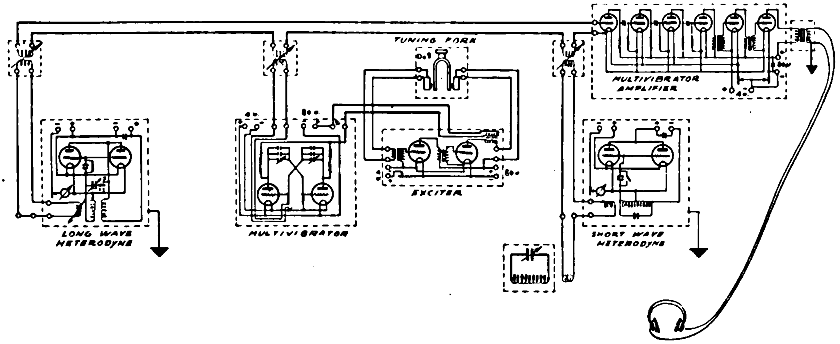

Fig. 1—Circuit Used in Measuring the Frequencies of Distant Radio Stations.

The multivibrator fundamental frequency is synchronized by means of a tuning fork giving approximately 1024 cycles per second. The tuning fork is operated and maintained in oscillation by the Eccles triode circuit, in the plate circuit of which is inserted the primary of a transformer whose secondary is in series in the multivibrator plate circuit (see Fig. 1).

Using this arrangement, when the multivibrator is approximately tuned to the tuning fork frequency a synchronizing effect takes place over a rather broad zone in the neighborhood of the tuning fork frequency. Thus the difficulty of a continuous control of the multivibrator frequency during the measurements is completely avoided.

Two heterodyne generators complete the setup. Fig. 2 illustrates the arrangement of the measuring apparatus.

Normally only one heterodyne generator is used. When it is necessary to standardize the shorter waves of the above-mentioned

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scale of frequencies the other generator is adjusted to the harmonics of the first one. Both heterodynes can be used for determining the order of the harmonics. For instance, the 96th harmonic of the multivibrator can be more easily found as the 8th harmonic of the 12th, or the 6th harmonic of the 16th, than by picking it up directly. Moreover, uncertainty is removed when there is the possibility of determining a particular harmonic by two different methods, especially when there is no possibility of obtaining adjacent harmonics, as in the example above. After the harmonics are identified, the calibration of the wavemeter is easily obtained by measuring with it

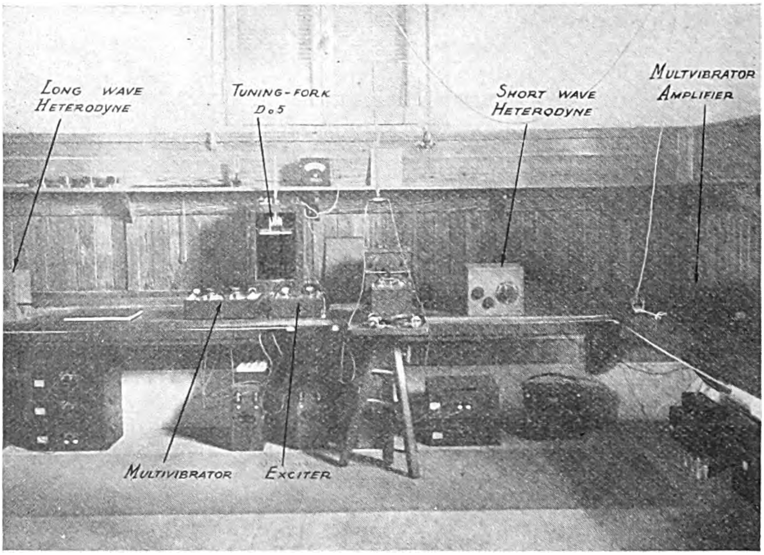

Fig. 2—Interior View of the Radio-Frequency and Calibratiom Set.

the frequencies of the electron-tube generators, successively adjusted to the different harmonics.

The measurement of a transmitting station frequency is obtained, first by synchronizing one of the generators on the unknown frequency, by means of a non-oscillating receiver, and then measuring that generator frequency with the calibrated wavemeter. If a very high precision is required the heterodyne oscillator may be used as a frequency meter. At the time of the measurement the multivibrator is adjusted to frequencies very near those to be measured and the corresponding reading of the heterodyne condenser observed. By successive interpolation the value of the frequency is obtained,

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which corresponds to the condenser setting for which the condition of zero beat exists between the heterodyne and the incoming signal. Every possible precaution should be taken during this procedure for maintaining the frequency of the heterodyne constant.

The piezo-electric crystals, now widely used in radio, can also be considered as high-frequency tuning forks, having a low temperature coefficient and sufficient stability to permit their use in the construction of very good secondary standards of frequency. The calibration of a piezo-electric crystal can be obtained by connecting it in the usual circuit, and then measuring its frequency by means of a calibrated electron tube generator. The same procedure described above for the measurement of a transmitting station frequency is used.

The tuning fork calibration, which is undoubtedly the principal problem connected with this kind of measurement, has required the development of a rapid and easy method, capable of giving satisfactory precision. This is obtainable in the usual electrotechnical laboratory practice.

First, it is necessary to consider how far it is possible to rely on the constancy of the tuning fork frequency.

It is known that for a tuning fork the number of vibrations of the fundamental frequency is directly proportional to the prong thickness, inversely proportional to the square of its length, and directly proportional to the square root of the fraction E/D, in which E is the modulus of elasticity and D the density of the steel of the tuning fork. Temperature, which affects directly the modulus of elasticity, is the principal factor influencing the frequency of the tuning fork (frequency).

It is known that between N, which represents the tuning fork frequency at temperature t, and N_o, which is the same frequency at zero degrees C, the following relation exists:

$$N = N_o(1 - \alpha t - \beta t^2)$$

Normally, β is very small and t^2 can be neglected, but the coefficient α in the common steel tuning forks varies between 0.000088 and 0.00012. Consequently, in the ordinary tuning forks there are frequency variations due to temperature changes which may be in some cases of the order of several parts per thousand.

In the electrically-driven tuning forks, other factors besides the temperature have an effect on the frequency stability, but their effects are not marked. It is known that the tuning fork frequency decreases when the amplitude of the tuning fork vibrations increases. Con-

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sequently, it is clear that any variation in the excitation current (that is, the filament current of the vacuum tube), or of the capacity shunting the exciting coils, or of the load applied to the secondary of the transformer, is accompanied by corresponding variations in the tuning fork frequency. A very accurate study of these subjects has been made by Dr. D. Dye.[1]

As a result of the above considerations we reached the conclusion that in measurements requiring a high degree of precision (1 to 5 parts in 100,000) it is necessary to calibrate the tuning fork each time before the measurement and to maintain an absolute constancy of temperature. When less precision (5 parts in 10,000) is required a

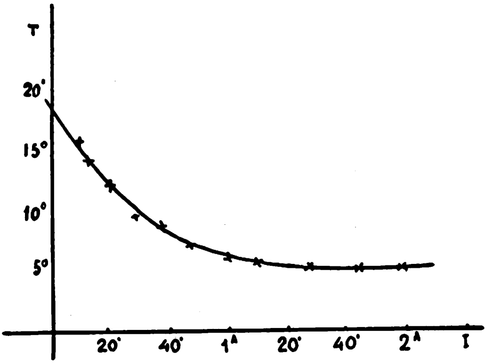

Fig. 3

former calibration can be used, but in this case the temperature coefficient must be known and the other conditions (relating to the tubes, plate current, etc.) must be exactly those which were used during the former calibration.

We believe that the exact determination of the tuning fork temperature, which lags behind the temperature of the surrounding air, constitutes the greatest difficulty of this measurement. For example, Prof. Gianfranceschi, who conducted investigations on the 870-cycle standard tuning fork of the Physical Institute of Rome, found a tuning fork cooling curve similar to that shown in Fig. 3. He has also found by calculation that in the case of the above mentioned standard tuning fork the coefficient α' of the equation of the curve

$$T = T_a - T_e \cdot e^{\alpha' T}$$

was given by:

$$\alpha' = 0.02339$$

[1] *Proc.* Roy. Soc. of London, p. 240, May, 1923.

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where T and T_a were, respectively, the tuning fork temperature at the instant t and the temperature of the air, and T_e represented the excess of the initial temperature of the tuning fork over that of the temperature of the air.

By the method we have developed we can obtain the calibration of a tuning fork very simply and quickly. In Fig. 4 is shown the stator winding of a small Siemens 100-toothed alternator of the induction type connected in series with the secondary of the transformer in the output circuit of the multivibrator amplifier. This alternator is driven by a d.c. motor which carries on the same shaft a small d.c.

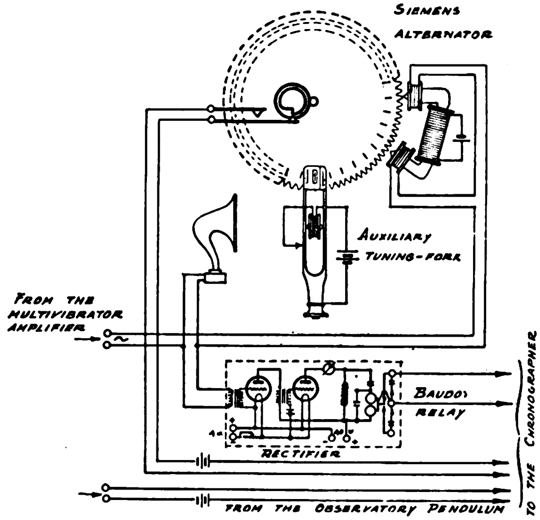

Fig. 4—Circuit Diagram of the Method Adopted for the Calibration of the Multivibrator Tuning Fork.

generator. In order to obtain a steadier speed of the revolving gear this generator can be loaded with a resistance coil of a suitable value. Moreover, a centrifugal speed regulator of the type used in the Hughes telegraphic apparatus is used on the revolving shaft for the same purpose. In order to obtain a frequency from the alternator as near as possible to the tuning fork frequency, a stroboscopic figure composed of 39 sectors is drawn on the alternator rotor and is observed through a small window, open at each semivibration, by an auxiliary electrically-driven tuning fork having a frequency of 200 cycles per second.

When the stroboscopic figure appears still as viewed through the

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small window, the number of revolutions per second n attained by the alternator is such that the following relation exists:

$$\frac{1''}{39\times n}=\frac{1''}{200\times 2}.$$

As the rotor has 100 teeth the corresponding frequency of the above conditions is given by:

$$f=\frac{400}{39}\times 100=1025.6$$

which is very near to the frequency of the tuning fork.

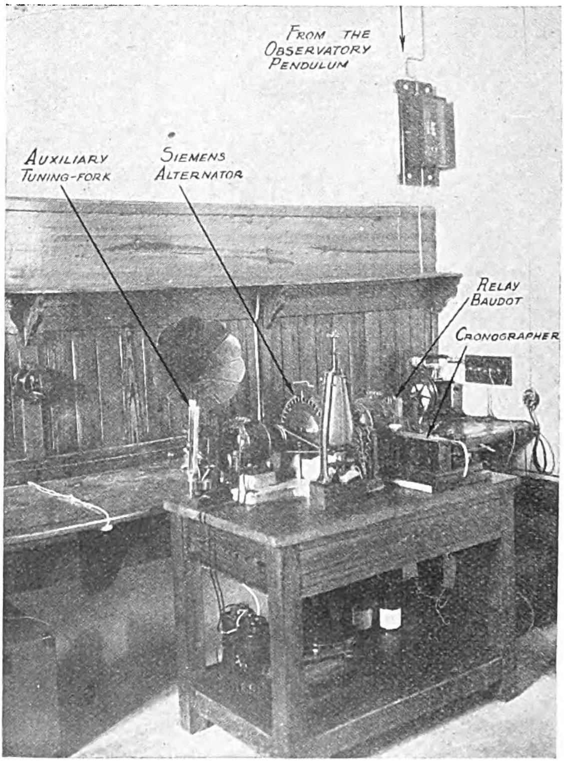

Fig. 5—Interior View of the Apparatus Used in the Tuning Fork Oscillation.

When the alternator frequency is close to the tuning fork frequency, beats will be heard in the loud speaker telephone in series with the amplifier output transformer of the multivibrator and the Siemens

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alternator. A suitable amplifier followed by a vacuum-tube detector shunts the loud speaker. In the plate circuit of the detector we have placed a very sensitive Baudot relay, having coils with 11,500 turns each, and a resistance of 5000 ohms per coil. This relay is operated at each beat.

The installation shown in Fig. 5 is completed by a three-pen recorder. One of the pens is actuated by the above-mentioned relay; the second is actuated by a contact which closes every 10 revolutions of the motor-alternator shaft, and the third is controlled by a contact on the standard pendulum in the Astronomic Observatory of the Campidoglio, which is received by means of a special line. Fig. 6 shows one of the records obtained. Dots (*a*) show the number of beats, the dashes (*b*) indicate the time intervals (a one-second dash each two seconds), and dots (*c*) furnish the number of tens of the alternator revolutions, that is, the number of thousand cycles of the alternator current.

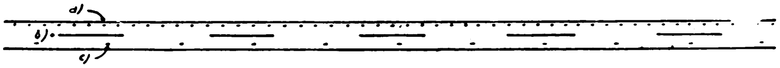

Fig. 6

We have found that the alternator speed need not be kept absolutely constant during the measurement, and that it is not necessary to observe the stroboscope continuously. Conditions are satisfactory for measurements when the difference between the alternator and the tuning fork frequencies remains the same as indicated by the note in the loud speaker. This condition is ordinarily obtained when beats are not too fast to be recorded. By means of the loud speaker it is possible to control at each instant the regular working of the beat recorder relay. If small variations of the alternator speed are required, they can be obtained by means of the d.c. generator loading resistance or directly by changing the mechanical regulating mechanism.

Evidently, if n is the number of the alternator cycles in a certain time t, and a is the number of beats recorded in the same time, the tuning fork frequency will be given by $n \pm a/t$. The sign to be applied at a can be obtained by observing the stroboscopic figure corresponding to zero beat. In the case of the Institute installation, at zero beat the stroboscopic figure appeared to revolve slowly in the direction of the shaft revolution. As the same figure normally had a greater speed in the same direction for the ordinary frequencies we came to the conclusion that the sign of a was negative.

Using a temperature controlled box (thermostat) we were able

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to obtain the calibration of the Institute tuning forks at various temperatures, and thus it was possible to obtain the temperature coefficient, which was 0.105 part in 1000. This is very near to that of 0.12 part in 1000 stated by the tuning fork manufacturer. Such a coefficient may be used for ordinary calibration in which case the tuning fork frequency can be calculated by means of the formula:

$$N = 1026.7 \times (1 - 0.000102t)$$

where t is the tuning fork temperature (Centigrade scale) that may be deduced from the air temperature by introducing proper corrections, and 1026.7 is the number of vibrations reduced to zero obtained from a long series of tests.

Using the above mentioned formula it is necessary to keep in mind that besides the uncertainty existing in the value of t, there is another source of error by assuming the quantity $N_o = 1026.7$ as a constant. This is because the supply current in the tuning fork circuit can change in the intervals between the various measurements, causing other differences of several parts in 100,000 from one test to another. This is especially true when the tests are made on different days. Consequently, when a high precision is required the tuning fork calibration must be repeated for each measurement.

Table I, which refers to measurements made on March 15, 1928 at a mean air temperature of 19 deg. C gives a clear demonstration of the precision of the present method.

TABLE I

CALIBRATION OF THE TUNING FORK

N	Date	Temperature	Seconds	Alternator cycles n	Beats Total number a	Tuning-fork cycles $N = n - a$	Tuning-fork frequency $f = \frac{n-a}{t}$	Mean frequency	Jumps from the mean %
1	4-3-928	19°C.	104	107.000	437	106.563	1,024,644	1,024,698	−0.054
2			566	583.435	3469	579.966	1,024,674		−0.024
3			326	335.610	1551	334.059	1,024,720		+0.022
4			488	502.733	2665	500.068	1,024,720		+0.022
5			1366	1,408.716	8951	1,399.765	1,024,720		+0.022

The principal advantages of the method described are the following:

(1) It permits the tuning fork calibration both before and after the various series of measurements, that is, under conditions which are practically the same as those existing during the use of the tuning fork as an exciter of the multivibrator.

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(2) It avoids the necessity of obtaining an absolutely constant speed of the alternator.

In order to demonstrate the method followed during the measurements and the degree of precision obtainable, we give the following two examples:

(1) The calibration of a piezo-electric crystal (Table II).

(2) The measurement of the frequency emitted by a tube transmitting set of the Rome (Torrenova) high power radio installation (Table III).

TABLE II

Calibration of the Piezo-electric Crystal No. 1

March 15th, 1928

$t=14$ deg. C. Measured tuning fork frequency: $f=1025.20$

Number of the observations	Readings of the heterodyne condenser for the harmonics		Reading of the heterodyne condenser at zero beats with the crystal frequency	Interpolation frequency	
	71st	70th			
Means of 5 observ.	822.96	987.98	840.3	71st harmonic 70th harmonic Crystal frequency	72789.75 71764.65 72682.10

March 18th, 1928

$t=13$ deg. C. Measured tuning fork frequency: $f=1025.42$

Means of 7 observ.	807.76	973.70	827.36	71st harmonic 70th harmonic Crystal frequency	72804.8 71779.4 72683.6
Means of 16 observ.	808.41	973.79	827.76	71st harmonic 70th harmonic Crystal frequency	72804.8 71779.4 72685.2

The general mean value of the three series of measurements made on March 15 and 18, 1928, gives 72684.5 as the crystal frequency. This value, at a temperature of about 13.5 deg. C differs by +3.44 parts in 100,000 from the value 72682 determined on December 12, 1927 by the Naval Radio Laboratory in Leghorn at a temperature of 14.8 deg. C, but using a different method.

It can be seen that the sign of the frequency difference is in accord with the temperature difference, so there is a perfect agreement between Rome and Leghorn measurements.

The control installation is completed by a series of five standard piezo-electric crystals calibrated as described above.

We believe that very quick determinations over a wide range of frequencies can be easily made by utilizing the oscillations and the harmonics produced by the above mentioned crystals together with an electron tube generator.

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TABLE III

MEASUREMENT OF THE FREQUENCY EMITTED BY TORRENOVA (ROMA) ELECTRON TUBE TRANSMITTER

March 17th, 1928

$t = 13$ deg. C. Measured tuning fork frequency: $f = 1025.22$

Number of the observations	Readings of the heterodyne condenser for the harmonics		Readings of the heterodyne condenser at zero beats with the Torrenova frequency	Interpolation frequency	Wavelength (for $V = 299.82.10^6$ m)
	98th	97th			
I Series					
1	906.0	967.8	Mean of	97th harmonic 99446.34	
2	906.0	967.8	8 observations:	98th harmonic 100171.06	
3	906.0	967.9	920	Torrenova freq. 100240.78	m 2290.8
Mean	906.0	967.8			
II Series					
1	906.7	968.0			
2	906.5	967.8	Mean of		
3	906.6	968.1	11 observations:	Torrenova freq. 100235.06	m 2991.2
			920.84		
Mean	906.6	968.0			
III Series					
1	907.0	968.3	Mean of		
2	906.9	968.2	15 observations:		
3	907.0	968.4	920.97	Torrenova freq. 100238.06	m 2991.0
Mean	906.97	968.3			
IV Series					
1	907.5	969.5	Mean of		
2	908.0	969.6	27 observations:		
3	907.8	969.5	920.76	Torrenova freq. 100207.5	m 2990.5
Mean	907.8	969.5			

By means of the methods and apparatus described above the Experimental Institute of Communications is now in a position to make frequency measurements of distant radio transmitting stations with accuracy and precision and to exercise, as required, the necessary control of the waves of radio transmitting stations.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

NOTE ON EARTH REFLECTION OF ULTRA SHORT RADIO WAVES*

By

E. H. LANGE

(School of Engineering, The Johns Hopkins University, Baltimore, Md.)

Summary—*In analytical investigations of the resultant pattern of electric intensity about an antenna, for the longer waves, the earth in the vicinity of the antenna has generally been considered as a perfect conductor. At sufficiently high frequencies, the reflected waves may differ considerably in magnitude and phase from perfectly reflected waves. The resultant distribution of electric intensity about an ultra short antenna depends upon the nature of the reflecting surface. Some computations and curves are given for the reflection coefficients and phase angles for various surface conditions, in conjunction with a horizontal ultra short antenna. Theoretical polar diagrams have been computed for various heights of horizontal antenna above the surface.*

CERTAIN experiments with ultra short radio waves have been described by H. Yagi,[1] in which measurements were made of the field distribution about short linear oscillators of 2.6 meters wavelength. The effects of earth reflection show to a marked degree in the pattern of electric intensities about a horizontal linear oscillator, in a plane perpendicular to the oscillator, and for various heights above the ground. The experimentally determined polar diagrams mentioned suggest the inadequacy of the assumption of perfect reflection for very short waves.

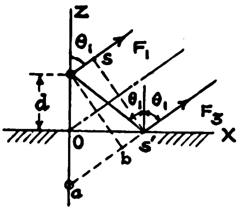

Fig. 1—Earth Reflection of Horizontal Electric Intensity.

The earth reflection from a horizontal linear oscillator is somewhat simpler to consider from the theoretical standpoint than that from a vertical oscillator[2]; also the experimental curves are more complete for the horizontal case, so that only the horizontal oscillator is here

* Dewey decimal classification: R113.6. Original manuscript received by the Institute, December 3, 1928.

[1] H. Yagi, "Beam Transmission of Ultra Short Waves," Proc. I.R.E. **16**, 715; June, 1928.

[2] Stuart Ballantine, "Review of Current Literature," Proc. I.R.E., **16**, 513; April, 1928; discussion of vertical dipol and bibliography.

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considered. The exact effect of the earth as a composite dielectric, with various degrees of conductivity and permeability, is necessarily very complex, and only the simplest assumption is considered in what follows for a non-magnetic layer having a certain amount of conductance and free charge capacitance.

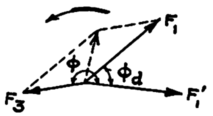

Fig. 2—Phase Relation of Direct and Reflected Intensity.

Polar diagrams for various heights of antenna above the earth have been computed for a plane perpendicular to the antenna, and are shown in Figs. 6 and 7. Points sufficiently far removed from the oscillator have been used in determining the directional intensity, so that the reflected and direct rays F_1 and F_3, Fig. 1, may be considered parallel in computing the phase difference at the wave front SS'. The resultant horizontal electric intensity thus depends upon the vector sum of the direct and reflected intensity, taking account of the magnitude and phase of each of the quantities at the wave front. Referring

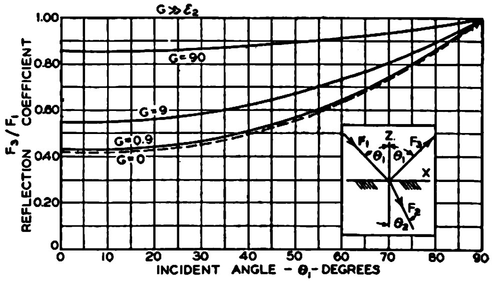

Fig. 3—Reflection Coefficients for Various Angles of Incidence, Resistivities, and Frequencies.

to Fig. 1 and Fig. 2, the excess path to the wave front for the reflected wave with reference to the direct wave is

$$ab = 2d \cos \theta_1$$

and the corresponding phase difference is

$$\phi_d = \frac{4\pi}{\lambda_1} d \cos \theta_1 \qquad (1)$$

A further phase change occurs at reflection, so that if F'_1 is the intensity before reflection, the direct wave at the wave front will be ϕ_d in

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advance of F_1', and the reflected intensity will have the phase angle ϕ with F_1' due to reflection. The resultant electric intensity is therefore the vector sum of F_1 and F_3.

For a perfectly reflecting surface the above computation is relatively simple, since the reflected and incident intensities are equal, and there is a phase change of 180 degrees on reflection. In this case the resultant phase difference between the direct and reflected waves at the wave front, and the resultant intensity, are determined by a simple expression. When, however, the reflecting layer has both conduction and displacement currents, both the magnitude and phase of the reflected ray vary with the angle of incidence.

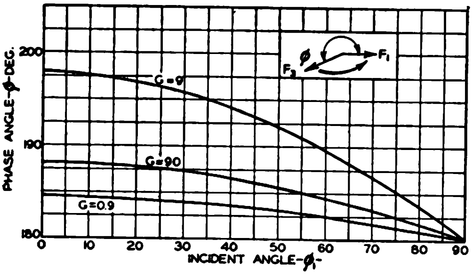

Fig. 4—Phase Angles for Various Angles of Incidence, Resistivities, and Frequencies.

From the following considerations, curves have been computed for the reflection coefficients and phase angles, for various angles of incidence and surface conditions. These curves are shown in Figs. 3 and 4. The electric intensities are f_x, f_y, and f_z, and the magnetic intensities h_x, h_y, and h_z.

The fundamental relations to be satisfied are the following Maxwell's equations in which $f_x = f_z = 0$, in accordance with the condition that the electric field is parallel with the earth's surface.

$$\frac{4\pi f_x}{\rho c} + \frac{\epsilon}{c}\dot{f}_x = \frac{\partial h_z}{\partial y} - \frac{\partial h_y}{\partial z} = 0 \quad (2)$$

$$\frac{4\pi f_y}{\rho c} + \frac{\epsilon}{c}\dot{f}_y = \frac{\partial h_x}{\partial z} - \frac{\partial h_z}{\partial x} \quad (3)$$

$$\frac{4\pi f_z}{\rho c} + \frac{\epsilon}{c}\dot{f}_z = \frac{\partial h_y}{\partial x} - \frac{\partial h_x}{\partial y} = 0 \quad (4)$$

$$-\frac{\mu}{c}\dot{h}_x = -\frac{\partial f_y}{\partial z} \quad (5)$$

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$$-\frac{\mu}{c}\dot{h}_y=\frac{\partial f_x}{\partial z}-\frac{\partial f_z}{\partial x}=0 \tag{6}$$

$$-\frac{\mu}{c}\dot{h}_z=\frac{\partial f_y}{\partial x}. \tag{7}$$

Therefore, the magnetic intensity $h_y=0$. (8)

Solutions of these equations are of the form

$$f=Fe^{-\alpha r}\times e^{j(\omega t-\beta r)} \tag{9}$$

where α and β are constants, $\omega=2\pi n$, and r the distance in the X-Z plane to any point from the earth's surface, along the beam.

$$r=x\sin\theta+z\cos\theta \tag{10}$$

Using subscripts 1, 2, and 3 to designate the incident, refracted and reflected beams, respectively, and noting that the value of α_1 for air is

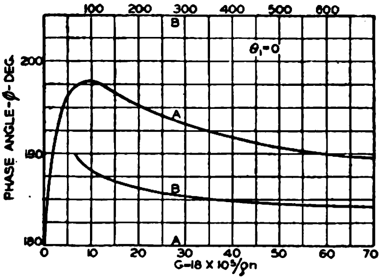

Fig. 5—Phase Angles at Normal Incidence for Various Resistivities and Frequencies.

negligible in comparison with α_2 for the earth, the electric intensities are

$$f_{y1}=F_{y1}e^{j(\omega t-\beta_1 r_1)} \tag{11}$$

$$f_{y3}=F_{y3}e^{j(\omega t+\beta_3 r_3)} \tag{12}$$

$$f_{y2}=F_{y2}e^{-\alpha_2 r_2}\times e^{j(\omega t-\beta_2 r_2)} \tag{13}$$

Considering the permeabilities $\mu_1=\mu_2=1$ for the air and the earth, also $\epsilon_1=1$ for the air, (3), (5), (7), and (9)

give
$$\frac{\omega^2}{c^2}\left(\frac{4\pi}{\rho\omega}j-\epsilon_2\right)=(\alpha_2+j\beta_2)^2 \tag{14}$$

for the earth, and
$$\frac{\omega}{c}=\beta_1=\beta_3 \tag{15}$$

for the air. From (5), (11), (12), and (13) the tangential magnetic intensities are:

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$$h_{x1} = -f_{y1}\frac{c\beta_1}{\omega}\cos\theta_1 \tag{16}$$

$$h_{x3} = f_{y3}\frac{c\beta_3}{\omega}\cos\theta_3 \tag{17}$$

$$h_{x2} = -f_{y2}\frac{cj(\alpha_2+j\beta_2)}{\omega}\cos\theta_2 \tag{18}$$

Equality of tangential components at the earth-air boundary requires that for any point at the boundary and for all instants of time,

$$f_{y1}+f_{y3}=f_{y2} \tag{19}$$

and

$$h_{x1}+h_{x3}=h_{x2} \tag{20}$$

From (19) and the values of (11), (12), (13), at the earth-air boundary, $\sin\theta_1 = -\sin\theta_3$ (21), that is, the angle of incidence equals the angle of reflection, and

$$\frac{\sin\theta_2}{\sin\theta_1} = \frac{j\beta_1}{(\alpha_2+j\beta_2)} \tag{22}$$

also

$$F_{y1}+F_{y3}=F_{y2} \tag{23}$$

and from (16), (17), (18), and (20)

$$F_{y1}-F_{y3}=F_{y2}\frac{j(\alpha_2+j\beta_2)}{\beta_1}\,\frac{\cos\theta_2}{\cos\theta_1} \tag{24}$$

Put

$$\frac{j(\alpha_2+j\beta_2)}{\beta_1}\,\frac{\cos\theta_2}{\cos\theta_1} = p+jq. \tag{25}$$

From (23) and (24) the vector ratio of the reflected electric intensity to the incident electric intensity is

$$\frac{F_{y3}}{F_{y1}} = \frac{1-(p+iq)}{1+(p+jq)}.$$

The ratio of the magnitudes is

$$\frac{F_{y3}}{F_{y1}} = \frac{\sqrt{(1-p)^2+q^2}}{\sqrt{(1+p)^2+q^2}} \tag{26}$$

and the phase angle ϕ between F_{y1} and F_{y3} is

$$\phi = \tan^{-1}\frac{-q}{1-p} - \tan^{-1}\frac{q}{1+p} = \tan^{-1}\frac{2q}{p^2+q^2-1}. \tag{27}$$

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The values of p and q for use in determining the amount and phase of the reflected electric intensity in terms of the incident electric intensity are determined from (14), (15), (22), and (25), and involve

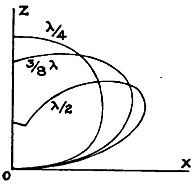

Fig. 6—Electric Intensities for Heights d of Antenna Above the Earth's Surface.

only the dielectric constant ϵ_2, the frequency n, the resistivity of the earth ρ, and the angle of incidence θ_1.

Thus, $$(p+jq)^2=\left(\epsilon_2-\frac{4\pi}{\rho\omega}j\right)\sec^2\theta_1-\tan^2\theta_1$$

Put $$b=\epsilon_2\sec^2\theta_1-\tan^2\theta_1$$

and $$g=\frac{4\pi}{\rho\omega}\sec^2\theta_1=G\cdot\sec^2\theta_1. \tag{28}$$

Then $$p=\sqrt{\frac{b+\sqrt{b^2+g^2}}{2}} \tag{29}$$

and $$q=\sqrt{\frac{-b+\sqrt{b^2+g^2}}{2}}. \tag{30}$$

The value of ρ in (28) is in electrostatic units. It will be more convenient to use ρ in ohms per centimeter cube, and the frequency

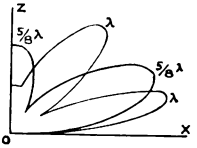

Fig. 7—Electric Intensities for Heights d of Antenna Above the Earth's Surface.

in megacycles per second. Therefore, with ρ = ohms per cm cube, and n = megacycles per sec.

$$G=\frac{18\times10^5}{\rho n}. \tag{31}$$

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As an instance of the magnitude of G, it may be noted that for $n = 100$ megacycles per sec. (i.e., 3 meters wavelength) and $\rho = 18 \times 10^3$ ohms per cm cube, $G = 1$. Thus, taking the value of $\epsilon_2 = 6$, the values of b and g which determine for normal incidence ($\theta_1 = 0$) the coefficient of reflection and phase angle are $b = 6$ and $g = 1$.

The curves of Figs. 3, 4, and 5 have been computed from (26) and (27), taking $\epsilon_2 = 6$, and the values of G ranging from zero to very large values compared with ϵ_2. For normal incidence, the phase angle ϕ due to reflection is greatest for a value of G in the vicinity of 9, when $\epsilon_2 = 6$. For any value of b, and for sufficiently small values of n, i.e., for the longer waves, G becomes large in comparison with ϵ_2, and the earth surface may be considered a good reflecting surface.

The polar diagrams have been computed for $G = 9$. In general form, there is good agreement with the experimentally observed distributions shown by Yagi, taking into account the fact that the observations were made at such proximity to the oscillator that the converging direct and reflected rays in some instances included an angle sufficient to affect the phase computation.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 4 *April, 1929*

Discussion on
PRINCIPLES OF GRID-LEAK GRID-CONDENSER DETECTION*

(F. E. Terman)

W. J. Polydoroff[1]: The theory and methods of grid-leak grid-condenser detection as developed by Professor Terman are very valuable to radio engineers, as they deal for the first time with numerical values of the circuit constants and actual results expected.

The phenomenon of detection has been well known to us for many years, and the values computed by Professor Terman are closely in accordance with those being used for several years. Unfortunately the detector tube requirements are considerably greater than that tube can meet because of increased sensitiveness and fidelity of present-day radio receivers. The range of sensitivity of a receiver if measured on the input of the detector circuit may be determined as 0.005 volt for very weak signals (distance reception) up to 10 volts on loud local signals.

Obviously present-day detector tubes can hardly take care of efficient rectification within such large limits. If the detector is adjusted to weak signal reception (maximum sensitivity), and therefore has high values of grid-leak resistance and small capacity, loud signals of local stations are very easily distorted and high notes distortion already existing due to sideband cut-off is accentuated still more. Besides, the detector is easily overloaded by strong signals producing grid polarizing voltage and general distortion occurs. As a result a usual precaution applied to most of the receivers is that *care should be taken not to overload the detector tube.*

A compromise is reached on values of a grid leak to be about 1 megohm and grid condenser of 250 $\mu\mu$f.

The anode bent rectification in this respect is considerably more elastic within the range of signal strength and introduces less distortion, but lacks efficiency and therefore reduces overall sensitivity of a receiver.

In my belief a two-element rectifier with substantially linear rectification would be a better solution for detection, especially when it is added to the set and the detector tube converted to act as a radio-frequency amplifier to make up for the loss of the detector's amplification. A sensitive and robust two-electrode tube can be designed to act as a rectifier in itself. I had some experience with similar detectors in the past and the only objection to them is that they do not amplify as they detect. The cost of an extra inexpensive tube would be justified for the increase of working range and better fidelity. The requirements of heating energy are so small that this tube can be operated from *B* supply of all electric sets saving the greater part of troubles due to a.c. hum encountered in their design.

However, at present we have to meet the situation with the tubes we have on hand and obtain the best results. The investigation of Professor Terman shows:

"Detector is less sensitive to high modulated frequencies, high notes are discriminated against,"

* Proc. I.R.E., 16, 1384; October, 1928.

[1] Johnson-Williamson Laboratories, Chicago, Ill.

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"Grid-leak resistance affects operating grid potential,"

"Change of grid potential is simply the voltage developed across the grid-leak condenser impedance."

These quotations are taken to prove that detector characteristics are *different* at *different frequencies* and at *different incoming voltages,* and therefore for the best results detector constants should be adjustable.

In the old days when the number of controls on a panel was considered more as an asset than a liability, receivers had as many as ten variable controls, among them adjustable grid-leak and adjustable grid-condenser. Nowadays, with a tendency to simplify the operation they are out of question.

However, I would like to show an effective way to control volume, especially adaptable to the receivers using a.c. tubes. We all know that the old methods of volume control by adjusting filament rheostats or plate supply cannot satisfactorily be applied to a.c. tubes, as a very small deviation in operating conditions of these tubes will inevitably produce objectionable hum. Consequently, it is usual to control input energy to the antenna circuit; this is especially convenient when aperiodic antenna and coupling tube are employed. For the receivers not neutralized and working under oscillating conditions this method is dangerous, as the tendency to oscillate increases when signal is weakened. There are a hundred points in the circuit where a resistance can be inserted to attenuate the incoming signal, in every case either affecting selectivity or shifting the resonance point. Control of the volume on the audio side is very favorable, but does not take care of detector overload.

A few months ago while trying to design an efficient volume control for a.c. operated loop receiver I encountered some difficulties, and after many comparisons came to a conclusion that a very practical method of volume control having distortionless properties is to use old variable grid leak for volume control.

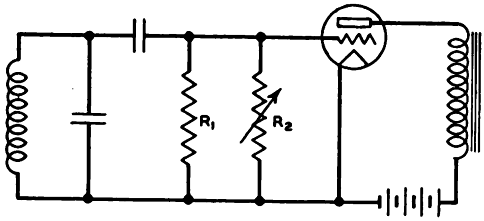

Fig. 1—The Circuit.
$R_1 = 2\omega$
$R_2 = 500{,}000 - 15{,}000$

A 2-megohm resistance is inserted in the usual way and in parallel with it a variable resistance 15,000–500,000 ohms. This variable resistance passes very small currents and therefore any plain carbon type rheostat, as used for plate supply control, can be inserted. When the resistance is in "off" position total resistance value is 2 megohms, and the detector is adjusted to maximum sensitiveness for weak signals. When 500,000 ohms are inserted in parallel it does not cause any noticeable decrease of reasonably loud signals, and for the local extra-loud signal the rheostat can be advanced to a desired volume and a signal can be brought to zero volume when total resistance is 15,000–50,000 ohms. These figures apply to any standard detector tube a.c. or d.c. type.

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The system, very simple but involving several modes of operation, performs the following functions:

(1) Adjusts detector for maximum sensitiveness with undistorted rectification for any given signal strength,
(2) Regulates output volume of radio receiver from maximum undistorted signal to zero without changing characteristics of other tubes,
(3) Improves fidelity of signal when volume is tuned down,
(4) Keeps r.f. amplification all the time at its full efficiency,
(5) Broadens the selectivity of the detector tuned stage.

First four points score to the advantage of this system; the last one is objectionable. The broadening effect occurs when variable resistance values are 50,000 ohms or less, in which case the tuned circuit is short-circuited by an impedance:

$$Z = \sqrt{\left(\frac{Rr_f}{R+r_f}\right)^2 + \frac{1}{C^2\omega^2}}$$

where R is variable resistance, r_f is grid-filament resistance, C is capacity of grid-condenser. According to Professor Terman's theory, the most important action which occurs in this scheme of control is the change of grid potential due to rectified current. When the signal is too great the rectified grid current changes the operated potential of the detector, which in turn causes overloading.

To avoid this, automatic means can be devised which would shift the operating grid potential of the detector in accordance with the magnitude of the carrier current receiver. This scheme is now under development and apparently offers a solution preventing detector overload. The volume control of the receiver in this case can be made in the audio parts of the apparatus, which, of course, offers no difficulties.

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BOOK REVIEW

Données Numériques de Radioélectricité. EDITED BY R. MESNY. Gauthier-Villars et Cie., Paris; McGraw-Hill Book Co., Inc., New York, 1928; pp. vii+26, quarto; price bound, 30 francs, unbound, 15 francs.

The book before us is that portion of Volume VI of the "Tables Annuelles de Constantes et Données Numériques" which is devoted to data on radio. Here, for the first time in the history of the Annual Tables, space has been found for a considerable amount of attention to various matters pertaining to radio communication. Hitherto the editors have not felt that radio data had attained a sufficient degree of precision to warrant fuller inclusion.

The usefulness to radio engineers of the Section under discussion is, of course, seriously limited by the fact that it includes only data that appeared in the course of two years, and that since then five years have elapsed. Nevertheless, the years 1923–24 brought forth so considerable an amount of material that its compilation from many widely scattered sources in the present volume is well worth while.

First of all we find several pages on numerical data and characteristic curves of tubes[1] and on electron emission from filaments. Then follow a few data on detector tubes and on magnetic permeability at high frequency. Wave propagation receives considerable attention, particularly Austin's measurements of intensity of signals from various stations. Results on direction finding are summarized, together with data on wave antennas and other forms of aerial.

Among other subjects treated are atmospheric disturbances, miscellaneous data on electrical oscillations, and the tables prepared by Grover and by Curtis and Sparks for the calculation of self-inductance. It is a striking tribute to the onward march of the vacuum tube that among all these data there is no word concerning arcs, sparks or contact detectors.

It is gratifying that the Annual Tables are now giving more space to the contributions which radio is making. While recognizing fully the great amount of labor and time required in selecting and editing such data as these, still we venture to hope that Professor Mesny will find it possible in future editions to bring the material more nearly up to date.

The section on radio can be obtained separately bound in board covers. Workers in the field will find the publication a very useful addition to their libraries.

W. G. CADY†

[1] Chiefly various types of G. E. tubes from articles in the PROC. I.R.E.

† Department of Physics, Wesleyan University, Middletown, Conn.

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MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The various articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R000. Radio Communication

R007 The problem of international distribution of broadcast wavelengths: Proposals of the Polish Broadcasting Co. *Experimental Wireless and Wireless Engineer* (London), 6, pp. 3–8; January, 1929.
(Deals with European situation.)

R010 Zenneck, J. The importance of radiotelegraphy in science. Proc. I. R. E., 17, pp. 89–114; January, 1929.
(A history of the progress of the art is given.)

R020 Hund, A. Hochfrequenz-Messtechnik. (High-frequency measurements). (book). 2nd edition, 1928. Published by J. Springer, Berlin. Noted in *Experimental Wireless* and *Wireless Engineer* (London), 5, p. 22; January, 1929.
(Review of book.)

R090 Important events in radio—peaks in the waves of wireless progress—1827 to 1928. Radio Service Bulletin, No. 141; pp. 16–26; December, 1928.

R100. Radio Principles

R113 Barfield, R. H. and Munro, G. H. The attenuation of wireless waves over towns. *Experimental Wireless and Wireless Engineer* (London), 6, pp. 31–37; January, 1929.
(Experimental and theoretical discussion of attenuation of radio waves over towns and suburbs. The absorption over towns is mostly due to buildings, vertical metal conductors, etc. and the absorption due to receiving stations plays only a small part, while over suburbs the absorption is mostly due to receiving stations.)

R113 Quäck, E. and Mögel, H. Hörbarkeitsgrenzen und gungstigste Verkehrszeiten bei Hurzwellen auf den einzelnen Überseelinien. (Audibility limits and most favorable communication directions for short waves over certain transoceanic paths). *Elekt. Nach. Tech.*, 5, pp. 542–549; December, 1928.

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(The charts given refer to a range of 30000 to 7500 kc. For twenty-four hours of service, one day and one night frequency are usually sufficient. For certain directions, a third wave is needed. If the rays of the sun are perpendicular to certain portions of the transmission path while the entire path is exposed to daylight, the reception is rather poor.)

R113.1 Colwell, R. C. Fading curves and weather conditions. PROC. I. R. E., **17**, pp. 143–148; January, 1929.

(Sunset fading curves from KDKA made at Morgantown, W. Va. with a Shaw recorder during April and May, 1927 showed that the signal strength from KDKA during the night was sometimes below and sometimes above daylight strength. Apparent correlation with weather.)

R113.3 Duckert, P. Elektrische Schwingungen und ihre Grenzgebiete drahtlose Telegraphie—Über Fehlweisung der Funkpeilung in Abhängigkeit von Wetterlage. (Electric waves and their limited transmission: On direction shift due to weather). *Zeits. für Tech. Phys.*, **9**, pp. 466–469; No. 12, 1928.

(Experiments carried on during several years (day measurements) show that the power received from a transmitter passes over a curved instead of a straight horizontal path so that the direction finder may be in error by several degrees. This is true when certain weather conditions exist, so called atmospheric border layers. During the night the direction deviations are also accompanied by fading.)

R113.4 Echos von Hertzachen Wellen (Echoes from electromagnetic waves). *Elek. Nach. Tech.* **5**, p. 488; December, 1928.

(Gives the different suppositions which may explain the echo effects. One explanation (Lassen and Försterling) assumes that the rays travel along different paths in the ionized layer. The second explanation (Störmer) assumes reflections or refractions on the envelope of electron pockets of very great dimensions. The third explanation (B. van der Pol, Jr.) assumes that the group velocity of the waves through the ionized layer can be slowed down considerably for certain conditions in the layer. The fourth explanation has reference to echoes corresponding to 0.005 to 0.08 second intervals, such as observed by Taylor and Young, Hoag and Andrew, and Quäck and Mögel. They are due to reflection from a layer 1500 km above the earth, from the polar nightlight zone, or from the sunset shadow wall.)

R113.4 Försterling, K. Über die Ausbreitung kurzer elektromagnetischer Wellen in der Heavisideschicht. (On the propagation of short electromagnetic waves in the Heaviside layer). *Elek. Nach. Tech.* **5**, pp. 530–542; December, 1928.

(General discussion on the refraction and absorption of short waves (less than 60 meters). Assumes that the ionization is due to ultraviolet rays from the sun. The article also indicates that there can be three layers from which the atmospheric ray may descend.)

R113.4 Appleton, E. V. Some notes on wireless methods of investigating the electrical structure of the upper atmosphere. *Proc. Physical Soc.* (London), **41**, pp. 43–59; December 15, 1928.

(Comparison of the wavelength change method, angle of incidence method, and the group retardation method for finding the effective height of the Kennelly-Heaviside layer. If the effect of the earth's magnetic field is neglected, the methods measure the same equivalent height which is greater than the maximum height reached by the atmospheric ray. A method for investigating the ionic concentration is given. This is partly based on an assumption by Pedersen for which the angle of incidence at the ground is greater than 30 deg.)

R113.5 Wright, C. S. Radio communication and magnetic disturbances. *Nature* (London), **122**, p. 961; December 22, 1928.

(Note on the effect of magnetic disturbances on radio reception for frequencies greater than 150 kc for crystal reception.)

R113.5 Chapman, S. The ultra-violet light of the sun as the origin of aurorae and magnetic storms (letter). *Nature* (London), **122**, p. 921; December 3, 1928.

(Refers to the letter by H. B. Maris and E. O. Hulburt (*Nature*, **24**, 1924) on their theory of aurorae and magnetic storms. This theory assumes that occasional sudden blasts of ultra-violet light produce the aurorae and magnetic storms. Chapman expects to offer another theory of the magnetic disturbances, which will assume the cause due to a neutral ionized stream (F. A. Linderman's original suggestion) which will avoid some of the inconsistencies of the Maris and Hulburt theory.)

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R114 Kenrick, G. W. The analysis of irregular motions with applications to the energy frequency spectrum of static and telegraph signals. *Philosophical Magazine* (London), 7, pp. 176–196; January, 1929.

(The analysis shows that the energy contained within a frequency band of given width due to an atmospheric disturbance produced by a random sequence of pulses varies directly with the square of the wavelength. A numerical example shows that the energy due to pulses of the order of 10^{-3} second in duration and sharply rising exponential form can produce appreciable fields (since proportional to the square roots of the energies) on the longer wavelengths.)

R124 von Ardenne, M. Einige Messungen uber die Hochfrequenzspannungen an der Eingangsseite von Empfangern. (Some measurements on r.f. voltages induced in coil antennas). *Zeits. fur Hoch.*, **32**, pp. 199–202; December, 1928.

(An aperiodic high frequency amplifier of known amplification inserted between the coil antenna and a tube voltmeter in order to bring the sensitiveness of the tube voltmeter to the proper value. Applications to field intensity measurements on broadcast stations are given.)

R125.1 Smith-Rose, R. L. The reversibility of radio direction finders and local error at rotating loop beacon. *Jour.* I. E. E. (London), **67**, pp. 149–156; January, 1929.

(It is shown that transmission from a rotating loop beacon being received on an open antenna produces about the same kind of errors as those experienced when the antenna is employed for transmitting to a direction finder. Methods are given for eliminating local errors.)

R125.6 Meissner, A. and Rothe, H. On the determination of the optimum radiation angle for horizontal antennas. Proc. I. R. E., **17**, pp. 35–41; January, 1929.

(The most favorable radiation angle for 15 and 20-meter wavelengths was determined using horizontal multiple antennas in connection with a parabolic reflector. It was found that the most favorable radiation happened when it took place along the tangent of the surface of the earth.)

R132 Colebrook, F. M. A generalized analysis of the triode valve equivalent network. *Jour.* I. E. E., (London), **67**, pp. 157–169; January, 1929.

(Analytical treatment (graphical) of a vacuum tube amplifier. It is based on Miller's assumptions (Bureau of Standards, **15**, p. 367; 1919) with modifications due to Hartshorn (*Proc.* Physical Soc., London, **39**, p. 108, 1926–1927). One conclusion is that at very high radio frequencies the voltage amplification factor may apparently exceed the voltage factor of the tube if a pure inductive plate load is provided.)

R132 Forstmann, A. and Schramm, E. Ueber Maximalleistungen von Verstärkerröhren. (On the maximum output of electron tube amplifiers). *Zeits. für Hoch.*, **32**, pp. 195–199; December, 1928.

(Formulas are derived for optimum output of vacuum tubes which are loaded either by a resistance or by an inductance. It is shown that for a constant plate potential and a resistance load the ratio of external to internal resistance is smaller than unity.)

R132 Snavely, B. L. and Webb, J. S. Radio frequency amplifying circuits. Proc. I. R. E., **17**, pp. 118–126; January, 1929.

(An equation is developed showing relation between circuit constants and critical grid resistance in a resistance stabilized amplifier having a tuned-grid circuit and a pure inductance plate load.)

R133 Kohl, K. Über kurze ungedämpfte elektrische Wellen. (On short sustained electromagnetic waves). *Zeits. für Tech. Phys.*, **9**, pp. 472–473; No. 12, 1928.

(The author attributes the frequency of ultra short waves in the Barkhausen-Kurz circuit partially to the natural frequency of an ordinary oscillating circuit. The frequency is increased on account of the dielectric constant of the electron gas between the grid and the anode since this constant is smaller than unity. The frequency is therefore not due to a pure electronic oscillation.)

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R133 Edgeworth, E. Frequency variation of the triode oscillator (letter). *Philosophical Magazine* (London), 7, pp. 200–203; January, 1929.

(Further discussion D. F. Martyn's paper (*Phil. Mag.*, Nov. 1927) with reference to the paper by E. Edgeworth in *Jour.* I. E. E., Wireless Section, Jan. 6, 1926.)

R134 Terman, F. E. and Googin, T. M. Detection characteristics of three-element vacuum tubes. Proc. I. R. E., 17, pp. 149–160; January, 1929.

(Comparative tests of different tubes using grid detection.)

R144 Butterworth, S. The high frequency resistance of toroidal coils. *Experimental Wireless and Wireless Engr.* (London). 6, pp. 13–16; January, 1929

(A discussion of the toroid coil in comparison to a single layer solenoid. It is shown that although the toroid has the advantage of being astatic it has always more high frequency resistance than a properly designed single layer coil and that the best possible toroid has more than twice as great a high frequency resistance as an equally compact single layer coil.)

R148 van der Pol, B. Some remarks on ultra short wave broadcasting. *Experimental Wireless and Wireless Engr.* (London), 6, pp. 9–12; January, 1929.

(Theory of pure amplitude modulation showing that for ordinary short wave transmitters a frequency modulation due to the resistance of the oscillator occurs in addition and that this interference can be avoided by using a piezo-electric controlled oscillator.)

R200. Radio Measurements and Standardization

R210 Hitchcock, R. C. A direct reading radio frequency meter. Proc. I. R. E., 17, pp. 24–34; January, 1929.

(A new type of direct reading r. f. meter.)

R210 Pierce, G. W. Magnetostriction oscillators. Proc. I. R. E., 17, pp. 42–88; January, 1929.

(Use of magnetostriction to produce and control electrical and mechanical frequencies of oscillations in a range of frequencies extending from a few hundred cycles per second to more than three hundred thousand cycles per second. Methods of calibration of vibrators and their use in the calibration of frequency meters and otner data pertaining to such oscillators are given.)

R214 Gerth, F. and Rachow, H. Die Temperaturabhangigkeit der Frequenz des Quarzresonators. (The temperature dependence of the frequency of quartz resonators). *Elekt. Nach. Tech.*, 5, pp. 549–551; December, 1928.

(Measurements of the absolute temperature coefficient of a quartz plate gave a value of 60 parts in one million per degree Centigrade. This figure was found for a temperature variation of 10 to 25°C. The quartz plate was silvered and determination is independent of the crystal holder.)

R214 Wright, J. W. The piezo-electric crystal oscillator. Proc. I. R. E., 17, pp. 127–142; January, 1929.

(Theory and operation of the piezo oscillator.)

R220 Wilmotte, R. M. A quick and sensitive method of measuring condenser losses at radio frequencies. *Jour. Scientific Instruments* (London), 5, pp. 369–377; December, 1928.

(The substitution method for measuring the effective condenser resistance has been improved by employing proper screening. Curves for several air condensers are given. The method can be used from up to approximately 6000 kc.)

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R220 Weihe, W. Die Messung von Kapazitäten mit dem Ueberlagerungsverfahren. (The measurement of capacities with the heterodyne principle). *Zeits. für Hoch.*, **32**, pp. 185–194; December, 1928.

(A beat note method using the silent zone is employed for measuring the effective grid-filament capacity for different external anode resistances (for 2000 to 10,000 kc.)

R300. Radio Apparatus and Equipment

R334 Westman, H. P. UV–861, a screen-grid tube for the high power amateur transmitter. *QST*, **13**, pp. 41–43; February, 1929.

(Characteristics of tube.)

R341 Pike, O. W. and Maser, H. T. A new type of rectifier tube for amateur use. *QST*, **13**, pp. 20–22; February, 1929.

(Description of UX-866 rectifier tube known as the hot cathode mercury vapor rectifier. Characteristics of tube are given.)

R351 Lampkin, G. F. An auxiliary frequency control for r.f. oscillators. Proc. I. R. E., **17**, pp. 115–117; January, 1929.

(A method is described for varying the frequency of an oscillator in small amounts by the use of a control which operates on the normally fixed element in the oscillating circuit.)

R386 Plebanski, J. Filtering antennas and filter-valve circuits. Proc. I. R. E., **17**, pp. 161–73; January, 1929.

(Describes methods of coupling together many circuits or antennas giving them simultaneous excitation from the same source of power. The purpose of these arrangements is the construction of practical filter circuits (filtering antennas) giving square topped resonance curves with good efficiency.)

R400. Radio Communication Systems

R401 Wollner, E. Die Fernsprechverbindung zwischen Europa und Amerika. (Long wave communication between Europe and America). *Elek. Nach. Tech.*, **5**, pp. 489–522; December, 1928.

(A review of transoceanic radio communication with long waves (corresponding to a range from 17 to 60 kc).)

R470 Bodie, C. A. and Curtis, R. C. The transmission of high frequency currents for communication over existing power networks. *Journal* A. I. E. E., **48**, pp. 37–41; January, 1929.

(Tuned choke coils are used to isolate the communication channel from the remainder of the power system. Such a system gives improved quality of speech, reduction in the noise level, freedom from variation in signal due to switching.)

R500. Applications of Radio

R526.4 The radio altimeter. *Science and Invention*, pp. 952–53; February, 1929.

(General Electric Company altimeter designed by E. F. W. Alexanderson.)

R536 Gleason, C. S. How radio prospecting takes the gamble out of mining. *Radio News*, **10**, pp. 716–719; February, 1929.

(Methods by which geologists study the distribution of minerals under ground.)

R592 Moullin, E. B. Radiotelegraphy and radiotelephony. *Jour.* I. E. E., (London), **67**, pp. 170-176; January, 1929.

(Review of many phases of radio. It includes the achievements in beam communication, electron tubes, transatlantic telephony, action of the Kennelly-Heaviside layer and other miscellaneous branches in radio.)

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R800. Non-Radio Subjects

537.65 Vigoureaux, P. Development of formulae for the constants of the equivalent electrical circuit of a quartz resonator in terms of the elastic and piezo-electric constants. *Philosophical Magazine* (London), 6, pp. 1140–1153; December, 1928.

(The equivalent electric constants of a quartz resonator are derived for the case of a long bar using Lamb's equation and the equivalence between the electrical and mechanical vibrating systems.)

621.385 Herd, J. F. The transmission unit and its application to radio measurements. *Experimental Wireless and Wireless Engr.*(London), 6, pp. 17–22; January, 1929.

(Brings out the merit of the TU and compares the same with the Napier unit used in some countries.

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PROCEEDINGS
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Institute of Radio Engineers

Forthcoming Meetings

CLEVELAND SECTION
Cleveland, Ohio, May 3, 1929

DETROIT SECTION
Detroit, Michigan, May 17, 1929

LOS ANGELES SECTION
Los Angeles, Calif., May 20, 1929

PHILADELPHIA SECTION
Philadelphia, Pa., May 24, 1929

PITTSBURGH SECTION
Pittsburgh, Pa., May 21, 1929

ROCHESTER SECTION
Rochester, N. Y., May 10, 1929

SAN FRANCISCO SECTION
San Francisco, Calif., May 17, 1929

TORONTO SECTION
Toronto, Ont., Canada, May 3, 1929

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PROCEEDINGS OF
The Institute of Radio Engineers

Volume 17 | May, 1929 | Number 5

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CARL DREHER   W. WILSON

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The Institute of Radio Engineers

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Year Books for 1926, 1927, and 1928, containing general information, the Constitution and By-Laws, catalog of membership etc., are priced at seventy-five cents per copy per year.

Contributors to the PROCEEDINGS are referred to the following page for suggestions as to approved methods of preparing manuscripts for publication in the PROCEEDINGS.

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Entered as second class matter at the Post Office at Menasha, Wisconsin.

Acceptance for mailing at special rate of postage provided for in the Act of February 28, 1925, embodied in paragraph 4, Section 412, P. L. and R. Authorized October 26, 1927.

Published monthly by

THE INSTITUTE OF RADIO ENGINEERS, INC.

Publication office, 450–454 Ahnaip Street, Menasha, Wis.

BUSINESS, EDITORIAL, AND ADVERTISING OFFICES,

33 West 39th St., New York, N. Y.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume, page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer divisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be 3/16 in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportion.

Abbreviations—Write a.c. and d.c., kc, μf, $\mu\mu$f, emf, mh, μh, henries, abscissas, antennas. Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper

Publication of Paper

Disposition—All manuscripts should be addressed to the Institute of Radio Engineers, 33 West 39th Street, New York City. They will be examined by the Committee on Meetings and Papers and by the Editor. Authors are advised as promptly as possible of the action taken, usually within one month.

Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

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MELVILLE EASTHAM
Treasurer of the Institute, 1927–1929

Melville Eastham was born in Oregon City, Oregon, on June 26, 1885.

For two years he was employed by a local electric light and street railway company. In 1905 he came to New York City and was chief engineer of Willyoung and Gibson, instrument manufacturers.

In 1906, with J. Emery Clapp, Mr. Eastham founded the Clapp-Eastham Company, one of the pioneer manufacturers of radio equipment and accessories. In 1915 with several other persons Mr. Eastham organized the General Radio Company. Since that time he has been President of the company. He is the designer of many instruments used in radio-frequency measurements and holds a number of patents on them.

Mr. Eastham has served on a number of the Committees of the Institute throughout his long active association with the Institute. He has been on the Board of Direction continuously since 1922. In 1927 he was elected Treasurer of the Institute, in which capacity he has continued to serve.

Mr. Eastham was elected to Associate membership in the Institute in 1913, was transferred to the Member grade in 1913, and to the Fellow grade in 1925. He is a member of the American Institute of Electrical Engineers, the American Physical Society, and the Optical Society of America.

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April Meeting of the Board of Directors

A meeting of the Board of Direction of the Institute was held on April 3rd in the office of the Institute in New York City. The following Board members were present: Ralph Bown, (acting chairman), Melville Eastham, treasurer; John M. Clayton, secretary; Arthur Batcheller, Walter G. Cady, R. A. Heising, Lewis M. Hull, C. M. Jansky, Jr., R. H. Manson, and R. H. Marriott.

One hundred and forty-four Associate members and fourteen Junior members were elected.

By unanimous vote of the Board, it was decided that the Institute annual awards for 1929 are to be as follows:

Institute Medal of Honor to Professor G. W. Pierce, of Harvard University.

Morris Liebmann Memorial Prize to Professor E. V. Appleton, of King's College, London.

Both of these awards were presented at the Banquet of the Fourth Annual Convention in Washington.

The Board authorized the appointment of a preliminary committee to investigate the possibility of the Institute's sponsoring the publication of a complete bibliography on radio.

Standard Frequency Transmissions by the Bureau of Standards

The schedule of standard frequency transmissions by the Bureau of Standards for the months of April and July (inclusive) appear below. This schedule includes many of the border frequencies between services as set forth in the allocation of the International Radio Convention of Washington which went into effect January 1, 1929. The signals are transmitted from the Bureau's station WWV, Washington, D. C. They can be heard and utilized by stations equipped for continuous-wave reception at distances up to 1,000 miles from the transmitting station.

The transmissions are by continuous-wave radiotelegraphy. The modulation which was previously on these signals has been eliminated. A complete frequency transmission includes a "general call" and "standard frequency" signal, and "announcements." The "general call" is given at the beginning of the 8-minute period and continues

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for about 2 minutes. This includes a statement of the frequency. The "standard frequency signal" is a series of very long dashes with the call letter (WWV) intervening. This signal continues for about 4 minutes. The "announcements" are on the same frequency as the "standard frequency signal" just transmitted and contain a statement of the frequency. An announcement of the next frequency to be transmitted is then given. There is then a 4-minute interval while the transmitting set is adjusted for the next frequency.

Information on how to receive and utilize the signals is given in Bureau of Standards Letter Circular No. 171, which may be obtained by applying to the Bureau of Standards, Washington, D. C. Even though only a few frequency points are received, persons can obtain as complete a frequency meter calibration as desired by the method of generator harmonics, information on which is given in the letter circular. The schedule of standard frequency signals is as follows:

RADIO TRANSMISSIONS OF STANDARD FREQUENCY; SCHEDULE OF FREQUENCIES IN KILOCYCLES

Eastern Standard Time	April 22	May 20	June 20	July 22
10:00–10:08 P.M.	4000	125	550	1500
10:12–10:20	4500	150	600	1700
10:24–10:32	5000	200	700	2000
10:36–10:44	5500	250	800	2300
10:48–10:56	6000	300	1000	2700
11:00–11:08	6500	375	1200	3100
11:12–11:20	7000	450	1400	3500
11:24–11:32	7300	550	1500	4000

Proposed Lehigh Valley Section

The second organization meeting of the proposed Lehigh Valley section was held March 8th in the Science Building of Muhlenberg College, Allentown, Pa. Carlton F. Maylott, temporary chairman, presided. Twenty-three members and guests attended the meeting.

The paper, "Recent Developments in Superheterodyne Receivers," by G. L. Beers and W. L. Carlson, was presented by Carlton F. Maylott.

Temporary officers were elected as follows: Carlton F. Maylott, chairman; Francis J. Hardner, vice-chairman; Fred. W Clauss, secretary-treasurer; Byron H. Eckert, chairman Meetings and Papers Committee; Byron H. Eckert, chairman Membership Committee; A. J. D. Haines, chairman, Publicity Committee.

Proposed Kansas City Section

Members of the Institute residing in the vicinity of Kansas City, Missouri, have expressed a desire to organize a Kansas City section of the Institute. E. A. McDowell, E. W. Hodge, and W. G.Wheat

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have been instrumental in the preliminary organization plans. An organization meeting was held in the Alladin Hotel, Kansas City, Missouri, on April 4th. E. A. McDowell presided.

W. J. MacDonald, Assistant U. S. Radio Inspector of that district, spoke on the work of the Radio Division of the Department of Commerce.

Institute Meetings

Boston Section

The Boston section held a meeting in the Cruft Laboratory, Harvard University, Cambridge, Mass., on March 14th. G. W. Pierce, chairman of the section, presided and ninety-eight members and guests attended.

Two papers were presented; one by C. F. Cairns entitled "Voltage Regulators," and another by G.W. Pierce entitled "High-Frequency Sounds." Messrs. Lamson, Putnam, Murray, McElroy, and Tyzzer participated in the discussion which followed.

Buffalo-Niagara Section

On March 29th the Buffalo-Niagara section held a meeting at the University of Buffalo, Buffalo, N. Y. L. C. F. Horle, chairman, presided. Sixty members and guests attended the meeting.

Lewis M. Hull, of the Radio Frequency Laboratories, presented a paper on "Recent Developments in Aircraft Radio."

Messrs. Eichman, Horle, Hector, and Lidbury participated in the discussion which followed.

Chicago Section

On March 8th a meeting of the Chicago section was held in the Engineering Hall of the Engineering Building, Chicago, Ill., and was presided over by the chairman, H. E. Kranz. The meeting was attended by seven hundred members and guests.

The paper, "The Kyle Condenser Reproducer," was presented by V. Ford Greaves. The speaker presented a mathematical study of the condenser reproducer, and showed the effect which is obtained in the Kyle speaker; where the heavy back plate is corrugated, an insulator between the dielectrics is used with an extremely thin conductor on the front, and a rolling action then takes place at the edges of the supporting ridges. Methods of connection of the condenser speaker to various types of circuits were also discussed, including data on the proper matching of impedances as well as the optimum resistance capacity ratio to secure a balanced reproduction. Numerous models of the

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speaker units, which are flat plates 8 x 12 inches, were shown in various stages of manufacture, as well as complete devices in various types of cabinets, screens, and other types of support. A large speaker having an equivalent area of 64 square feet was used for the major part of the demonstration, and feeding it from an amplifier using four type 250 tubes in the output circuit, the quality was everything that could be desired. Tuning forks of 56, 112, 1024, 4096 and 8192 cycles all came through clearly when struck and held before the microphone.

Messrs. Sabine, Kranz, Kyle, and others participated in the discussion which followed.

On April 5th, at the Electric Club, Chicago, Ill., the Chicago section held their regular meeting, presided over by H. E. Kranz, chairman of the section. Fifty-eight members and guests attended the meeting.

A. E. Shaw, of the Ryerson Laboratory of the University of Chicago, presented a paper, "Cold Cathode Rectification." This paper is published in this issue of the PROCEEDINGS.

Messrs. Morrison, Kranz, Eppstein, and Miller participated in the discussion which followed.

CLEVELAND SECTION

In the Case School of Applied Science, Cleveland, Ohio, the Cleveland section held a meeting on March 29th. Bruce W. David, chairman of the section, presided. There were fifty-three members and guests in attendance.

R. G. Ransom, of the American Telephone and Telegraph Company, presented a paper, "Chain Broadcasting—A New Art." A brief history of chain broadcasting was given beginning with the "wire transmission" of a ceremony at the Arlington Cemetery in 1922. Problems in the transmission of high quality programs such as distortion due to unequal attenuation were discussed and characteristic curves shown. The paper was illustrated by lantern slides.

LOS ANGELES SECTION

The Los Angeles section held a meeting March 18th in the Elite Cafe, Los Angeles, California. T. F. McDonough, chairman of the section, presided. Sixty-five members and guests attended.

Mr. Carter and Mr. Mitchell, of the Radio Corporation of America, presented a moving picture film, "Manmade Miracles," showing the production of R. C. A. Radiotrons.

Bert Fox presented a paper, "Recent Developments in Superheterodyne Receivers," by G. L. Beers and W. L. Carlson.

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K. Dalton demonstrated a power amplifier audio receiver using a fifty-watt output tube and feeding nine dynamic speakers. The rectifier supplying power to the entire system is an arrangement of four 281-type radiotrons in a bridge-type circuit.

New Orleans Section

The New Orleans section held a meeting March 27th in the Louisiana Engineering Building, New Orleans, La. J. N. Du Treil, vice-chairman of the section, presided. There were fifty-one members and guests in attendance.

J. N. Du Treil, Radio Inspector of the United States Department of Commerce, presented a paper, "Crystal Oscillators." The paper covered various crystals that exhibited piezo-electric effects, methods of grinding and cutting crystals, precaution being observed in cutting them. The various circuits in which they might be used and the advantages of each were also explained. Mr. Du Treil exhibited samples of crystal blanks, finished crystals, and complete General Radio crystal oscillator for broadcast station use.

Messrs. Val Jensen, Jones, Gardberg, and Mackie participated in the discussion which followed.

New York Meeting

The monthly New York meeting of the Institute was held on April 3rd in the Engineering Societies Building, 33 West 39th Street, New York City. In the absence of President Taylor, R. H. Marriott presided.

The papers presented at the meeting comprised a series on the general subject of frequency measurement. These papers were as follows:

"The Routine Measurement of the Operating Frequencies of Broadcast Stations," by Henry L. Bogardus and Charles T. Manning, presented by Mr. Bogardus.

"The Testing of Piezo Oscillators for Broadcasting Stations," by E. L. Hall.

"The Precision Measurement of Time," by Alfred L. Loomis.

"A High Precision Standard of Frequency," by W. A. Marrison.

"Observations on Modes of Vibration and Temperature Coefficients of Quartz Crystal Plates," by F. R. Lack.

"A Convenient Method for Referring Secondary Frequency Standards to a Standard Time Interval," by L. M. Hull and James K. Clapp, presented by Mr. Clapp.

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"Frequency Measurement," by S. Jimbo, presented by Walter G. Cady.

It is expected that these papers will be published in the PROCEEDINGS.

Four hundred and fifty members of the Institute and guests attended this meeting.

Philadelphia Section

On March 22nd, at the Franklin Institute, Philadelphia, Pa., the Philadelphia section held their regular meeting. The meeting was presided over by J. C. Van Horn, chairman, and was attended by twenty-five members and guests.

This was a business meeting and the following officers were elected, to serve until June 1930: J. C. Van Horn, chairman; C. Brown Hyatt, vice-chairman; John C. Mevius, secretary-treasurer.

An open discussion followed on "The Trend of Radio Receivers."

Rochester Section

The Rochester section held a meeting February 1st in the Sagamore Hotel, Rochester, New York. A. B. Chamberlain presided. There were seventy-eight members and guests in attendance. This was a joint meeting with the Rochester Section of the American Institute of Electrical Engineers.

Sedwick M. Wright, of the General Railway Signal Company of Rochester, was the speaker, having as his subject "Centralized Traffic Control."

On March 15th the Rochester section held a meeting in the Sagamore Hotel, Rochester, New York, presided over by A. B. Chamberlain, chairman of the section. This was a joint meeting with the Rochester Engineering Society and the local section of the American Institute of Electrical Engineers. Eighty-six members and guests attended.

Edward C. Jerman, of the Victor X-Ray Corporation of Chicago, Ill., was the speaker. His subject was, "Progress in X-Ray in Medicine, Research, and Industry."

San Francisco Section

On March 20th the San Francisco section held a meeting in the Hotel Bellevue, San Francisco, California. The meeting was presided over by Donald K. Lippincott, chairman of the section. Forty members and guests attended.

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Dr. F. E. Terman presented a paper, "The Possibilities and Limitations of Intelligence Transmission with a Limited Band of Frequencies." The paper told of the great increase in number of radio stations and how this increase makes necessary steps towards improving of conditions in order to accommodate them on the available frequency channels. Dr. Terman explained that broadcasting telephones require about 5,000 cycles of channel space, speech telephones require about 2,500 cycles, 60-line picture transmission about 15,000 cycles, and telegraph stations about 35 to 130 cycles, depending on speed of transmission.

The paper brought out the fact that more stations could be accommodated than possible at present by utilizing means now at our disposal, and stated that new developments were not necessary.

As remedies for congested conditions the speaker gave; first, better frequency control; secondly, multiple modulation; thirdly, frequency multiple systems; fourthly, assignment of frequency bands to radio companies, allowing them to crowd as many stations as possible into them.

In order to standardize frequencies a system was suggested to have a primary frequency standard station and several secondary stations to transmit control frequencies and to use these frequencies as master controls, multiplying to secure the desired frequency output.

A brief discussion by the members present followed the presentation of the paper.

Seattle Section

On February 26th, the Seattle section held their regular meeting in the Telephone Building, Seattle, Washington. W. A. Kleist, chairman of the section, presided. There were forty-eight members and guests in attendance.

W. A. Kleist presented an informal paper on "Sound Reproduction, Apparatus for Motion Picture Theatres." Some of the apparatus for this purpose was demonstrated.

Messrs. Libby, Tolmie, and Eastman participated in the discussion which followed.

The annual election of officers was held with the following results Austin V. Eastman, chairman; Abner R. Willson, secretary-treasurer.

The Seattle section held their regular meeting March 29th in the Philosophy Hall of the University of Washington. Austin V. Eastman, chairman, presided. Fifty-four members and guests attended the meeting.

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F. M. Reynolds, of the General Electric Company, presented a paper "Keyport 30-kw Tube Transmitter." The paper described the construction and operation of the 30-kw tube transmitter just installed for the Navy Department at Keyport, Washington. One of the new features of this transmitter is the use of mercury-vapor filled hot cathode rectifier tubes for plate current supply. The keying circuit cuts in and out grid bias voltage on the master oscillator circuit to start and stop these tubes oscillating. The amplifier uses two 20-kw tubes in parallel. The antenna current is about 100 amperes at 50 kc. The radiation resistance is about two ohms at this frequency. The transmitter may also be operated at 102 kc.

Messrs. Libby, Deardorf, Threlkeld, Kleist, and Willson participated in the discussion which followed.

Toronto Section

The Toronto section held a meeting March 13th in the Electrical Bldg., University of Toronto, Toronto, Canada. The meeting was presided over by A. M. Patience, chairman of the section, and one hundred members and guests attended. This was a joint meeting with the Rochester and Buffalo-Niagara sections of the Institute.

Irving Wolff, of the Radio Corporation of America, presented a paper, "Problems of Small Moving Iron Type Speakers."

Messrs. Bayly, Karker, Hector, and Pollock participated in the discussion which followed.

On April 3rd, the Toronto section held a meeting in the Electrical Bldg., University of Toronto. A. M. Patience, chairman, presided. Two hundred members and guests attended the meeting.

J. H. Vennes, of the Northern Electric Company, presented a paper on "Audible Motion Pictures."

Messrs. Patience, Pipe, Elliott, Bayly, and Burns participated in the discussion which followed.

Washington Section

In the Continental Hotel, Washington, D. C., on March 14th, the Washington section held a meeting. The meeting was presided over by F. P. Guthrie, chairman of the section. Seventy-two members and guests attended the meeting.

Captain Guy Hill, of the Engineering Division of the Federal Radio Commission, presented a paper on "The Re-allocation of Broadcast Frequencies."

Following the presentation of this paper, Commissioner La Font

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and Mr. Dellinger gave interesting talks on the activities of the Federal Radio Commission.

Messrs. Guthrie, Gunn, Hyland, Robinson, Brown, Pettsing, and Smith participated in the discussion which followed.

Committee Work

Committee on Admissions

A meeting of the Committee on Admissions was held in the office of the Institute at 9:30 A.M. on April 3rd. The following members were present: R. A. Heising, chairman; F. H. Kroger, C. M. Jansky, Jr., E. R. Shute, J. S. Smith, and A. F. Van Dyck.

The Committee considered twenty-two applications for transfer or election to the higher grades of membership in the Institute.

Committee on Sections

A meeting of the Committee on Sections was held on April 2nd. Those present were E. R. Shute, chairman; Austin Bailey, B. E. Shackelford, John M. Clayton, and C. J. Porter.

The Committee considered the revision of the Sections Manual of Organization and Operation.

The Committee outlined in detail the program for the Annual Meeting of Section Delegates at the Annual Convention in Washington, D. C.

Personal Mention

Lloyd J. Andres, formerly chief engineer, Musical Device Corporation of Chicago, has recently taken charge of the talking picture and sound projection equipment design and development for the Dramaphone Corporation of Chicago.

Lester D. Culley, recently of the Department of Physics, University of California, is now a receiving engineer of the Radio Corporation of America at Marshalls, Cal.

Edwin Mraz is now engineer with the Electrical Research Laboratories, Inc., of Chicago. Mr. Mraz was formerly chief engineer of the Greene, Brown Mfg. Co.

G. Edwin Stewart has recently become associated with the Paramount Famous Lasky Corporation of New York City as chief recording engineer, having resigned his position as maintenance engineer with the National Broadcasting Company of New York.

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C. L. Walker is now inspection engineer in the Radio Engineering Department of the American Bosch Magneto Corporation at Springfield, Mass.

H. F. Wareing, late instructor in radio with the University of Wisconsin, has recently become associated with the Universal Wireless Communication Company, Inc. in the position of engineer in charge of the Transmission Laboratory at Chicago.

J. H. Pressley, formerly a radio engineer with the Hazeltine Corporation of Hoboken, is now chief engineer of the U. S. Radio and Television Corporation at Marion, Ind.

Hollis S. Baird, until recently consulting engineer for the Automatic Radio Company of Boston, Mass., is now associated with the Short-Wave and Television Laboratory of Boston as chief engineer.

J. S. Douglas has recently become proprietor of the Peerless Radio Shop of Chicago.

Russell A. Cline has recently assumed connection with the Bell Telephone Company of Canada as assistant transmission engineer, Central Plant Department. Mr. Cline formerly was engaged as consulting engineer for Radio Station CKCL of Toronto.

Pierre Boucheron, until recently advertising and publicity manager of the Radio Corporation of America of New York City, is now sales manager of the Southern District of that organization at Atlanta, Ga.

Lyman T. Newell is now associated with the Airways Division of the Bureau of Lighthouses at Station WWX. Mr. Newell was formerly employed by the Radio Corporation of America as operator.

Paul R. Nachemson has recently become laboratory supervisor for the Brooklyn Radio Service Corporation, having left the Eagle Radio Company of Brooklyn where he formerly was employed as engineer.

Elmer B. Lyford was for some time an engineer with the Electrical Testing Laboratories of New York City and is now recording engineer with the RCA Photophone, Inc., of New York.

A. R. Hopkins, until recently associated with the Radio Laboratory of the Day Fan Electric Co. of Dayton, Ohio, is now in the employ of the Radio Corporation of America as radio engineer in the Technical and Test Department.

Oliver Wright, until recently a student at the University of Arizona, is now a technical employee of the American Telephone and Telegraph Company at its Lawrenceville, N. J. radio station.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED APRIL 3, 1929

Elected to the Associate grade

California	Los Angeles, 5308 Third Ave.	Howe, Franklin Joachino
	Los Angeles, 521 Amethyst St.	Oodrys, Arthur
	Palo Alto, Federal Telegraph Co.	Shermund, Ralph C.
	San Jose, 1148 Lincoln Ave.	Saine, Harry
	Stanford University, Box 1331	Skinner, Clifton Ross
Connecticut	Bridgeport, 68 Mead St.	Tillman, John Elmer
	Kensington, P.O. Box 276	Edgerly, A. H., Jr.
Dist of Columbia	Washington, Radio Section, Bureau of Standards	Arnold, Prescott N.
	Washington, 1705 Bay St., S.E.	Bower, Ward E.
	Bellevue, Naval Research Laboratory	Hyland, Lawrence A.
	Washington, 2301 Cathedral Ave., N.W.	Jackson, William E.
	Washington, 900 F St., N.W.	Kaplan, Jacob
	Washington, 2618-13th St. N.W.	Moulton, Theodore S.
	Washington, The Chastleton, 1701 16th St. N.W.	Newton, Jane Elizabeth
Illinois	Chicago, 5054 Agatite Ave.	Buesman, Robert Marshall
	Chicago, 6716 Parnell Ave.	Cubert, Joseph R.
	Chicago, 411 S. Ashland Blvd.	Kempf, Frank J.
	Chicago, 3250 North Crawford Ave.	Kincaid, Owen D.
	Chicago, 501 North Central Ave.	Knutson, Ben M.
	Chicago, 1706 N. Mayfield Ave.	Lamb, Harold A.
	Chicago, 3823 S. Kedzie Ave.	Patterson, Joseph A.
	Chicago, 3621 E. 107th St.	Siovic, John E.
	Chicago, 7728 Calumet Ave.	Van Sickel, R. E.
	Litchfield, Box 214	Hopper, C. L.
Indiana	Evansville, 112 Jackson Ave.	Norwood, Vandle Clarence
Iowa	Des Moines, 1538-31st St.	Hutchison, Sam T.
Kansas	Pittsburg, Pittsburg Radio Service	Forman, H. W., Jr.
Louisiana	New Orleans, 2625 Jefferson Ave.	Goldstein, Henry R.
Maine	Yarmouth, Greeley Road	Tamminen, Nestor
Massachusetts	Belmont, 98 Payson Road	Anthony, A. W., Jr.
	Boston, 472 Massachusetts Ave.	Nunez, F. J.
	Boston, Custom House, c/o Supervisor of Radio	Weston, Irving L.
	Cambridge, 22 Bigelow St.	Battison, Wallace A.
	Cambridge, General Radio Co., 30 State St.	Thiessen, Arthur E.
	Chatham, c/o Radio Corporation of America	Thing, K.
	North Attleboro, 9 Elm St.	Hale, Willis L.
	Revere, 62 Malden St.	Fletcher, Frederick R.
	Salem, 28½ Grove St.	Heffernan, S. K.
	Springfield, 62 Kimberly Ave.	Ferguson, George Well
	Stockbridge, Box 982	Harrington, George N.
Michigan	Alpena, 111 Tawas St.	Snyder, Leslie T.
	Ann Arbor, 1030 Church St.	Mathison, Gerald
	Detroit, 3028 Lothrop Ave.	Woehr, William
Minnesota	Minneapolis, 1521 University Ave., S.E.	Reed, Harry F.
Missouri	Bucklin	Holmlund, A. Earle
	Mendota, R.F.D.	Davis, Glen A.
	St. Louis, 4047 W. Pine St.	Humphreys, Irl W.
Nebraska	Falls City, 1801 Morton St.	Chesley, Arthur D.
New Hampshire	Claremont, 227 Main St.	Hodge, V. W.
New Jersey	Bloomfield, 25 Grace St.	Henry, T. J.
	Camden, 303 North 6th St.	Galvin, Robert
	Clayton, 326 Broad St.	Purvis, Charles G.
	Deal, Box 122	Shaw, Robert C.
	Glen Ridge, 129 Midland Ave.	Grabo, Irving C.
	Linden, 104 Luttgen Place	Angle, Gafford Brock
	New Brunswick, R.C.A. Radio Station	Bohman, Victor A.
	New Brunswick, 149 Hale St.	Lucas, Earle F.
	Stanhope, Box 261	Peterson, Arthur C.
New York	Astoria, L.I., 14-34 Grand Ave.	Ractliffe, Charles Lione
	Brooklyn, 392 Sackman St.	Berner, Aaron
	Brooklyn, 1121 Bedford Ave.	Dean, Leon W.
	Brooklyn, 506 Amboy St.	Epstein Reuben
	Brooklyn, 421 Chester St.	Haynes, Nat
	Brooklyn, 1300 New York Ave.	Herdman, Raymond C.
	Brooklyn, 1257 New York Ave.	Hesse, Henry Richard
	Brooklyn, 1859-62nd St.	Sass, Isidore
	Brooklyn, 55 Johnson St.	Skinker, Murray F.
	Buffalo, 191 Franklin St.	Bandetson, Harry
	Greenport, 829 Main St.	Larsen, Carl L.
	Jamestown, 72 Campbell Ave.	Beâumont, William Frederick
	New York City, 1050 Park Ave.	Cohn, Ralph I.
	New York City, 617 West 141st St.	Kahn, Morton B.
	New York City, 760 West End Ave.	Littman, Leon
	New York City, 160 West 100th St.	Lopez, Melchor, Jr.

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New York (con't)	New York City, 2740 Marion Ave., Bronx	Myers, Theobald
	New York City, 500 Riverside Drive	Sicari, Domenic
	New York City, 404 Riverside Drive	Turner, Eugene T., Jr.
	New York City, 781 East 182nd St., Bronx	Wheeler, George D.
	Richmond Hill, 8512 110th St.	Hudtwalker, William Theodore
	Riverhead, P.O. Box 982	Trevor, Bertram
	Rochester, 55 Hortense St.	Dwyer, Vincent J.
	Rochester, 207 Ave. C	Wiebach, William T.
	Rocky Point, c/o R.C.A.	Goldstine, Hallan E.
	Schenectady, 105 Seward Place	Clarke, Varro J.
	Schenectady, 422 Y.M.C.A. Bldg.	Lynn, L. H.
	Schenectady, 842 Union St.	Orr, Robert W.
	Whitestone, L. I.	Summers, Llewelyn L. B.
Ohio	Akron, 160 Fir St.	Ehrisman, Henry O.
	Ashtabula, R.D. No. 2	Andrus, Roy E.
	Cincinnati, 1212 Sassafras St.	Verkeley, B. M.
	Cincinnati, 3484 Vine St.	Wells, Martin M.
	Columbus, 34–18th Ave.	Auckerman, U. H.
	Columbus, Ohio State University, E.E. Dept.	Roetken, A. Allen
	Dayton, 43 Victor Ave.	Franzwa, Frederick J.
	Gambier, Kenyon College	Cottrell, Casper L.
	Lakewood, 2028 Waterbury Road	Hood, William A.
	Marion, 360 Silver St.	Ackerman, Francis R.
	Niles, 424 Allison St.	De Cola, Rinaldo
	Youngstown, 3630 Market St.	Pennock, P. L.
Oklahoma	Norman, 132 Page St.	Moffett, Le Roy, Jr.
	Oklahoma City, 1624 East Park Place	Pata, Yaromir J.
	Tulsa, Radio Station KVOO	Golder, Frank E.
Oregon	Portland, 605 Marguerite Ave. N.	Trumbull, A. F.
Pennsylvania	Allentown, 531 N. 7th St.	McGee, Michael J.
	Easton, 528 Centre St.	Messinger, Reuben R.
	Lancaster, 36 S. Lime St.	Russell, Walter L.
	Narberth, 403 N. Narberth Ave.	Bates, Clifford W.
	Philadelphia, 2030 E. Hazard St.	Martino, Alphonso E.
	South Tamaqua	Delp, Paul L.
	Upper Darby, 297 Springton Road	Lewis, Oliver I.
	Wilkinsburg, 901 South Ave.	Armstrong, Ralph W.
Rhode Island	Providence, 160 Cypress St.	Adams, Raymond R.
Tennessee	Memphis, 1961 Harbert St.	Brooks, Maurice W.
Texas	Dallas, 2603 Madera St.	Bennett, Porter T.
	Dallas, 1205½ Elm St.	Godard, L. G.
	Dallas, 5004 Goodwin Ave.	Melroy, Harry C.
	Fort Worth, 907 W. T. Waggoner Bldg.	Zeidlik, William J.
	Moody	McCauley, E. Ray, Jr.
Virginia	Alexandria, Box 107 R4	Meyers, Paul F.
	Marion	Cummings, G. N.
Washington	Seattle, 1833-13th Ave.	McAvoy, Edward J.
Wisconsin	Menasha, 526 Keyes St.	Peerenboom, Cyril A.
	Sheboygan, 922 Clara Ave.	Flentje, Le Roy G.
	Stoughton, 313 S. Academy St.	Turner, Russel S.
	Wausau, 108 Grand Ave.	Krueger, Otto J.
Canada	Montreal, Que., Northern Electric Co., Ltd., Keefer Bldg.	Cash, J. Allen
	Montreal, Que., 89 Laurier Ave. E.	Minorc, J. L.
	Montreal, Que., 4581 Sherbrook St. W.	Wilcox, B. B.
	St. Hyacinthe, Que., 7 Bourassa St.	Chagnon, Adolphe
Denmark	Copenhagen, Kastelsvej No. 3	Bertzow, Johannes Andreas
England	Chelmsford, Essex, 26 Queen's Road	O'Neill, R. F.
	Leigh-on-Sea, Essex, 106 Western Road	Richardson, F. C.
	London, N.W.6, 42 Fairhazel Gardens, Hampstead	Erdman, H.
	London, SW5. 12 Trebovir Road	Megaw, E.
	Parkstone, Dorset, St. Nicholas Castledene Road	White, John Walter
Holland	2 Dedelstraat den Haag	Bartelink, E. H. B.
Japan	Kumamoto, Kumamoto Broadcasting Station JOGK	Hirose, M.
	Kumamoto, Kumamoto Broadcasting Station	Morita, Kazuyoshi
	Tokyo, Teinshin-kanri-renshusho siba Park	Tei, Sasaki
	Kumamoto, c/o Shimizu-Hosojo	Kanayama, Toyosaku

Elected to the Junior Grade

California	Oakland, 627 Poirier St.	Ford, Charles Y.
Dist. of Columbia	Washington, Loomis Radio College	Carlson, O. D.
Illinois	Lake Forest, 1199 Edgewood Road	Baker, James Maurice
Indiana	Valparaiso, 405 E. Monroe St.	Ratts, Bruce H.
Iowa	Iowa City, c/o Kappa Eta Kappa, 728 Bowery St.	Stauffer, Ray Everett
Kansas	Manhattan, 413 N. 17th St.	Kipp, Aaron
Michigan	Battle Creek, 245 Lake Ave.	Fay, Lewis C.
Montana	Bozeman, 201 South Third	Pelton, George A.
New York	Brooklyn, 886 Putnam Ave.	Osterland, Edmund
	Brooklyn, 24 Bay 31 St.	Weinberg, Sidney
Ohio	Columbus, 75 West 10th Ave.	Conrad, Willard Oliver
Oklahoma	Stilwell	Neeley, Carl E.
Pennsylvania	Philadelphia, 653 North 10th St.	Huttenloch, Robert M.
Canada	Toronto, Ont., 10 Kew Beach Ave.	Hamox, Harold
England	Gloucester, 48 Weston Road	Carlyon, C.

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APPLICATIONS FOR MEMBERSHIP

Applications for election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to election of any of these applicants should communicate with the Secretary on or before June 1, 1929. These applicants will be considered by the Board of Direction at its June 5th meeting.

For Transfer to the Member grade

Dist. of Columbia	Washington, Federal Radio Comm.	Butman, Carl H.
	Washington, 1018 Douglas St., N. E.	Young, Leo C.
New Jersey	Newark, Brandes Products Corp., 200 Mt. Pleasant Ave.	Gerns, Wm. H.

For Election to the Member grade

Dist. of Columbia	Washington, 1821 Rigga Place, N. W.	Schrenk, M. H.
New York	New York City, c/o Imperial Japanese Navy, Inspectors Office, 1 Madison Ave.	Nakamura, T.
Czechoslovakia	Prague 11, u Karlovu 5,	Zacek, August
England	Cambridge, Sidney Sussex College	Ratcliffe, J. A.
Germany	Berlin-Wilmersdorf. Guntzelstr. Berlin	Watanabe, Yasusi

For Electoin to the Associate grade

Alabama	Mobile, c/o Adams Glass Co., Inc. 17-19-21 So. Royal	Heyen, Rudolph J.
California	Hollywood, 1123 Poinsettia Drive	Signer, John J.
	Inglewood, 826 65th St.	Green, Charles M.
	Los Angeles, 4846 Rosewood Avenue	Cooper, George R.
	Los Angeles, 1601 So. Hope St.	Holcomb, E. Russell
	Los Angeles, Western Air Express Inc., 117 W. 9th St.	Hoover, Herbert, Jr.
	Los Angeles, 316 N. Reno St.	Ludlum, William T.
	Los Angeles, 921 W. 6th St.	McGonagil, H. A.
	Los Angeles, 2812 So. Main	Meyer, R. H.
	Los Angeles, 1609 Hipoint St.	Walter, Edward F.
	San Francisco, 1266 McAllister St. Apt. 8	Slavin, Samuel B.
	San Francisco, 715 Bush Street	Stanton, James M.
Connecticut	Hartford, 235 Collins St.	Heris, George
	Hazardville, South St.	Thompson, Stanley W.
Dist. of Columbia	Bellevue, U. S. N. Lab.	Wortman, E. C.
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Part II

TECHNICAL PAPERS

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

DOUBLE AND MULTIPLE SIGNALS WITH SHORT WAVES*

By

E. Quäck and H. Mögel

(Transradio, Berlin, Germany)

Summary—*Multiple signals have been observed in a frequency range of 20,000 to 16,667 kc, or 15 to 18 m, only when the great circle between transmitter and receiver passes through the twilight zone, in which conditions are especially favorable for these frequencies.*

The conditions for various positions of great circles are discussed in detail, and the expected times of occurrence of double signals are given and compared with those measured. As the greatest angle which the plane of the circle must form with the twilight plane at the most favorable time of day and year, in order to receive multiple signals, the value $\phi \leqq 10$ *deg. is assigned.*

The propagation conditions in the twilight zone appear, at least for Berlin as receiving point, to be more favorable in the winter in the indirect sense, and in the summer in the direct. For reception from the west, this has been clearly confirmed for North and South America, while the direction of summer double signals from the east is not yet entirely determined. Next year, after the increase and improvement of transmitters, the observation of the lines favorable for double signals from Osaka, Mukden, and Manila will be possible.

Indirect double signals also appear in the range of 25,000 to 13,636 kc (12 to 22 m), outside of the twilight zone, when the indirect night path is in summer. Under these circumstances the same condition of the atmosphere prevails at night as in the daytime, perhaps even the same as at twilight.

Lower frequencies, between 12,000 and 6,667 kc (25 and 45 m) can also pass along the indirect path at night, when the solar radiation on the illuminated portion is not so intense. The amplitude received on the direct path is at this time just on the border of audibility.

An example cited from Baires shows that the indirect double signals can be supported or even amplified by reverse radiation from the transmitter.

The frequency ranges are given, in which the various double signals occur. Double signals have been found thus far between 25,000 and 6,667 kc (12 and 45 m).

The time intervals between main and double signals and between multiple signals themselves were obtained experimentally and by calculation, on the assumption that the velocity was the same as that of light in free space. It was found in all cases that the path was about 3.5 per cent longer than the actual distance over the surface of the earth. Observations extending over two years show within the limits of error no such variations in time intervals as those indicated in the results of Taylor and Young. Variations in time intervals between main and double signals could always be traced to an irregular beginning and cessation of the signal due to faulty keying.

Methods are indicated for screening off the indirect double signals by suitable antenna apparatus at the receiving end.

With the appearance of magnetic disturbances the double signals disappear more

* Dewey decimal classification: R113. Original manuscript received by the Institute, November 14, 1928. Translation received, January 11, 1929.

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completely the more the earth's field is disturbed and the more the angle between the direction of the earth's field and the transmission direction approaches a right angle.

A graphic representation is shown of amplitudes received from New York as a function of time of day and year, giving the amplitudes of double signals in per cent of the main signals; these results are published in somewhat different form, for the main lines of Transradio.[1]

The various great circles are shown graphically with data concerning the distribution of day, twilight, and night as a function of the time of year. The great circles Berlin-Rio-Mukden and Berlin-New York are indicated for January 1 and July 1. One can see at a glance the illumination of the direct and indirect paths, and can gain an idea of the possibility of communication, and particularly can determine in advance the times for double signals.

Regarding the work of A. H. Taylor and L. C. Young we should like to point out that we recently succeeded in ascertaining incontestably the existence of so-called short range echoes, which are discussed in the following supplement.

1. Introduction

SUPPLEMENTING previous publications concerning the occurrence of double and multiple signals on short waves,[2] there are presented here additional observations and a summary of the results to the present time.

All observations were made at the Geltow receiving station of the Transradio A.G. für drahtlosen Uebersee Verkehr. It must be mentioned that the routine receiving work prevented systematic experiments, so that the results to be reported were largely derived in the course of regular commercial operation. These are to serve as stimulation for further observations. In the meantime the interesting investigations of Taylor and Young[3] and of Eckersley[4] have been reported, the results of which will be discussed below.

The observation of double signals has, of course, a special value, since it undoubtedly extends our still meager positive knowledge concerning the transmission of short waves and the structure of the Heaviside layer. The observations made in Geltow and the results achieved can be considered as rather definite, since they extend over more than two years. Certain principles were set up at the very beginning, which have been further followed out, and have been so far very well confirmed. Nevertheless new phenomena are observed almost daily, the explanation of which will take time. Also as already

[1] *Elec. Nach. Tech.*

[2] E. Quäck, *Jahr.* **28**, p. 177, 1926, and *Elek. Nach. Tech.*, **4**, p. 74, 1927. See also *Jahr. der draht. Tel.*, **30**, No. 2, p. 147, 1927. For an account of these papers in English see Stuart Ballantine: "Review of Current Literature," PROC. I. R. E., **15**, 341; April, 1927, and **15**, 1065; December, 1927.

[3] A. H. Taylor and L. C. Young: Proc. I. R. E., **16**, 561; May, 1928.

[4] Eckersley: *Jour.* I. E. E., 1927, pp. 600–614.

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mentioned, the commercial receiving and the experimentation cannot be easily combined.

A description of the experimental methods used will first be given.

2. Apparatus and Precision

For the analysis of low-frequency currents coming from the receiver the Siemens loop oscillograph with three loops is used, which conveniently permits the photographic recording of continuous phenomena. The holder takes a strip of sensitive paper five meters (15 ft.) long, and the speed of the paper can be regulated from about 2 cm per second to 2 m per second. Only loops having a natural frequency of 10,000 cycles are used, in order to exclude loop effects, etc., and in

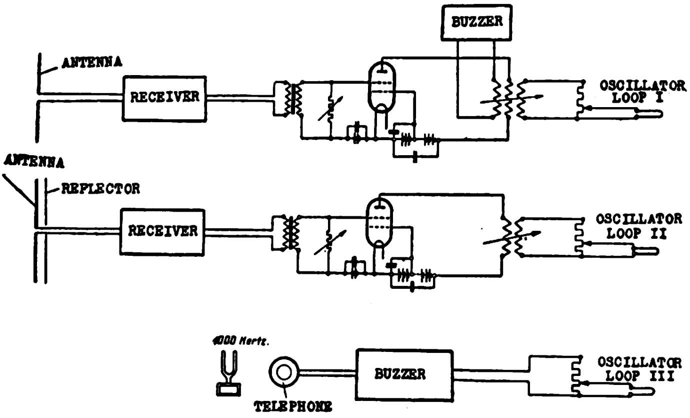

Fig. 1—Apparatus for recording of double signals (antenna, reflector, receiving set, audio generator, oscillograph loops).

order to record as exactly as possible the beginning of the signals or of oscillation groups of the order of 0.001 second. The requisite current of about 100 milliamperes for the loops is obtained by connection of one stage of low-frequency amplification with a Siemens OR tube. The amplitude of the loops can be varied by resistance and coupling.

For the determination of time a comparison frequency of 1000 cycles is also recorded, which is produced by a loosely coupled audio-frequency tube generator, which is occasionally tuned to a sonic fork. The precision of this arrangement (Fig. 1) is not required to be exceedingly high since other errors, amounting to 1 to 2 per cent, are present; for example, the beginning and cessation of signals often requires 1/1000 to 1/5000 sec., due to faulty keying. Since the quantities measured

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range from 1/10 to 1/30 sec. the precision is seldom greater than 1 per cent. Therefore, it is not possible with these means to detect and explain phenomena, the measured values of which lie within the order of magnitude of these errors.

3. Discussion of Oscillograms

First of all, a number of oscillograms of double and multiple signals will be shown, with the necessary explanations. Further discussion will follow in the succeeding sections.

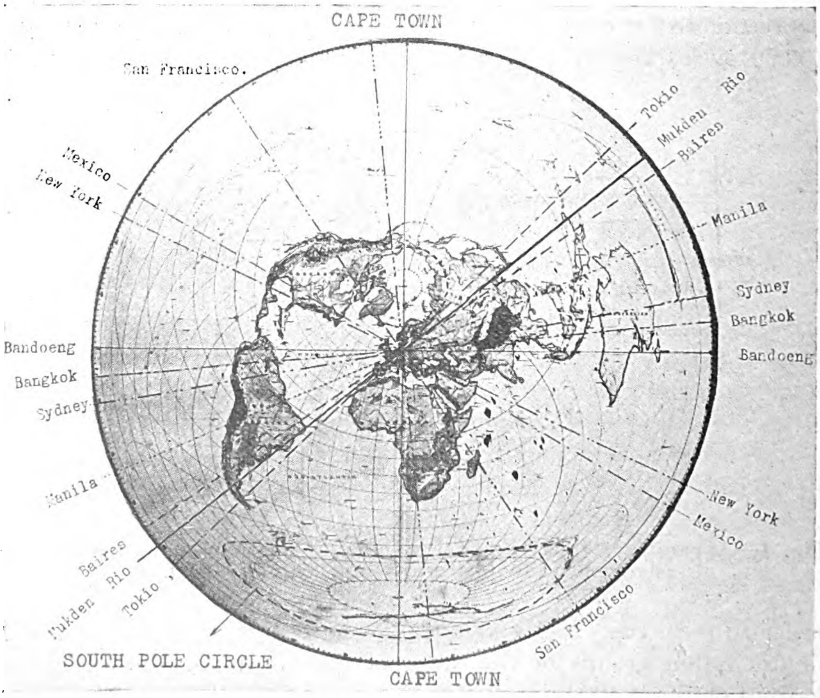

Fig. 2—Azimuthal projection of the earth, showing Geltow in the center as pole, the antipodal point being represented by the outer circle. The principal great circles appear as straight lines.

For the sake of clearer distinction we shall call those double and multiple signals which travel in the direction of the direct signal (that is, forward around the earth), "direct" double signals and "direct" multiple signals, respectively; those, however, which take the other course around the earth (that is, circulate in the reverse direction), "indirect" double signals and "indirect" multiple signals, respectively.

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The photographs are arranged according to the various directions (Fig. 2), and are further subdivided with regard to direct and indirect double signals. The double signals belonging to one main signal are designated by the same letter; the direct multiple signals, which pass around the earth several times in the same direction, are designated by superscripts, e.g., a^1, a^2, a^3, while for the indirect signals subscripts are used, as a_1, a_2, etc. Any signal indicated by a letter without index or subscript passes by the shortest route between transmitter and receiver.

I. South America

As examples of direct double signals the following are presented. Direct double signals from the short-wave transmitters in Rio* and Baires* appear in spring between 9 and 10 P.M., Central European Time, C.E.T. Gradually they grow stronger, reaching their maximum on the longest day. The directions toward Rio and Baires deviate only a few degrees (about 5 deg.) from each other, as shown by the map (Fig. 2), so that both can be considered as practically similar.

Fig. 3—Direct double signals from Rio, SPU (19,231 kc, or 15.6 m). April 11, 1928, 10:45 P.M., C.E.T. $\Delta t = 0.1385$ sec.

The record (Fig. 3) of April 11, 1928, shows weak double signals from station SPU ($f = 19{,}231$ kc, or $\lambda = 15.6$ m) in Rio; the difference in time, $\Delta t = 0.1385$ sec., corresponds to the earth's circumference. As a result of faulty keying a small insertion precedes every signal, which reappears exactly in the double signal and permits the difference in time to be determined more exactly than would otherwise be possible. The amplitude is very weak at this season, yet attains values of 30 per cent of the main signal. The following records, Figs. 4 and 5, of the same station SPU were made at about the time of the longest day, on June 26, 1927, about 9 P.M., C.E.T. At least five repetitions of the two short signals sent from Rio at an interval of about 1/25 second could be clearly traced here. In the upper record of Fig. 4, the signals a and b pass around the earth again in the same sense (a^1, b^1). In the lower exposure, Fig. 4, which was taken about $\frac{1}{3}$

*** Editorial note: These terms, "Rio" and "Baires," presumably refer to Rio de Janeiro, Brazil, and Buenos Aires, Argentina, respectively.**

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second later, the signals pass completely around the earth a second time in the direct sense (e.g., b^1, b^2), in addition to which also weak, indirect, double signals (a_1, b_1) can be traced. The differences in time amount in the first case (direct) to 0.1385 second, in the second case (indirect) to 0.069 second. Fig. 5, which was photographed a few minutes later, contains, after a further time difference of 0.138 sec.,

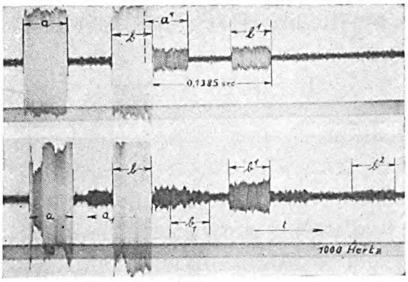

Fig. 4—Rio, SPU (19,231 kc, or 15.6 m), June 26, 1927, between 9 and 9:05 P.M., C.E.T. Triple signals b, b^1, b^2 in direct direction ($\Delta t = 0.1385$ sec.), and simple indirect signal b_1 ($\Delta t = 0.069$ sec.)

even a third weak direct signal, a^3, which accordingly has travelled over 130,000 km. The next oscillogram, Fig. 6, from station LP2 in Baires, shows direct double signals again with the normal difference in time $\Delta t = 0.138$ second. Fig. 7 shows a variation of the amplitude of station SPU taking place within $\frac{1}{4}$ second; this is of interest as showing that the fadings of the main and double signals bear no relation to each other.

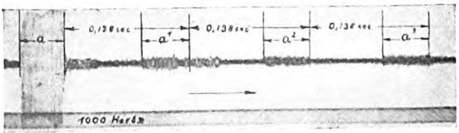

Fig. 5—Rio, SPU (19,231 kc, or 15.6 m), June 26, 1927, between 9 and 9:05 P.M., C.E.T. Four-fold direct signals (3 echoes). $\Delta t = 0.138$ sec.

There follow some pictures of indirect double signals. In the record, Fig. 8, from the Telefunken station LP3 in Baires, the main signals and indirect double signals overlap each other, so that the groups (Morse signals) appear longer than normal. The individual signals, which form the letter l (·—· ·), can, however, be easily read, while in the following picture, Fig. 9, from Rio SPU, this is no longer the case. The speed of sending is here so unfavorable that main and double signals merge completely. At any time, when both signals overlap,

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the amplitudes are added and can give the most varying pictures according to the amplitude and phase relations of main and double signals. The modulation of the transmitter must also be taken into consideration here.

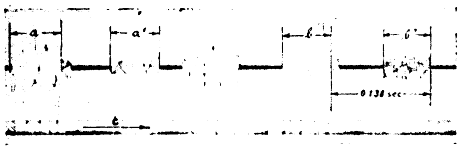

Fig. 6—**Baires, LP2 (20,690 kc, or 14.5 m), May 20, 1928, 9:30 P.M., C.E.T. Direct double signal, $\Delta t = 0.138$ sec.**

In Fig. 10, from the Baires station LP3, there appear besides the main signal a also two indirect double signals a_1 and a_2, a_1 with a time difference $\Delta t_1 = 0.056$ second after a and a_2 with $\Delta t_2 = 0.138$ second after a_1. In Figs. 11 and 12, a and b are simultaneous records from the

Fig. 7—**Rio, SPU (19,231 kc, or 15.6 m), May 31, 1927, 9:33 P.M., C.E.T. Fading of direct double signals (F) within 1/4 sec.**

Baires station LP3 ($f = 19{,}608$ kc, or $\lambda = 15.3$ m) and LP2 ($f = 20{,}690$ kc, or $\lambda = 14.5$ m) showing multiple indirect signals. In Fig. 11 (from LP3) the first indirect signal a_1 is stronger than the main signal a. Amplitude and phase happen to be such that the total amplitude

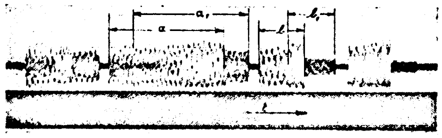

Fig. 8—**Baires, LP3 (19,608 kc, or 15.3 m), Feb. 18, 1928, 12:20 P.M., C.E.T. Indirect double signal, $\Delta t = 0.056$ sec.**

becomes greater during the overlapping of a and a_1. There appears also another signal a_2 which has passed around the earth again in the indirect sense with a time difference $\Delta t_2 = 0.138$ second with respect to a_1. On the same record (Fig. 11) in the case of LP2 the variation in the amplitudes of main and indirect signals within one second can be

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easily recognized. The same is the case in Fig. 11b, with LP3, where on the left side the main signal b has a much greater amplitude than the indirect double signal b_1, while on the right side the double signal c_1 is stronger than the main signal c. In Fig. 12b the main signal a and the indirect double signal a_1 from LP3 happen to have about the same amplitude, while the phase is shifted about 180 deg., so that the result-

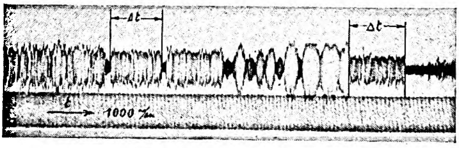

Fig. 9—Rio, SPU (19,231 kc, or 15.6 m), May 19, 1927, 3:45 P.M., C.E.T. Disturbance caused by indirect double signal. (Doppler effect.)

ing amplitude through overlapping becomes almost 0. There would accordingly seem to exist two separate groups of oscillations. The real length of the emitted signal is derived from the indirect round-the-world signal a_2, which with respect to a_1 has a time difference of $\Delta t_2 = 0.138$ sec. On the same record at the right below with LP2 the amplitude drops during the overlapping twice to 0, so that one receives the impression of beats or of a transmitter that is strongly modulated at about 20 cycles. (Perhaps Doppler effect). Furthermore, with LP2 there are no round-the-world signals to be found on this record.

II. North America

As a result of the rather unfavorable location of the great circle through New York (Rocky Point) and Berlin (Geltow) (see Sec. 4

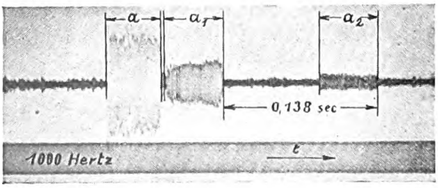

Fig. 10—Baires, LP3 (19,608 kc, or 15.3 m), Dec. 2, 1927, about 9.55 A.M., C.E.T. Indirect multiple signal. $\Delta t_1 = 0.056$ sec., $\Delta t_2 = 0.138$ sec.

and the map, Fig. 2) there appear here in general only double signals. Moreover, the times of occurrence of direct double signals are rather narrowly limited. Fig. 13 of July 13, 1928, 2 A.M., (C.E.T.), shows direct double signals of the North American station WTT ($f = 18{,}750$ kc, or $\lambda = 16$ m). In the upper record the double signals fall within the

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main signals; in the lower one they are separated from them. The double signals b_1 and c_1 have with respect to the main signals b and c a time-difference of 0.138 second. The disturbances happened to be very great here, so that the double signals stand out very indistinctly in the disturbed pattern.

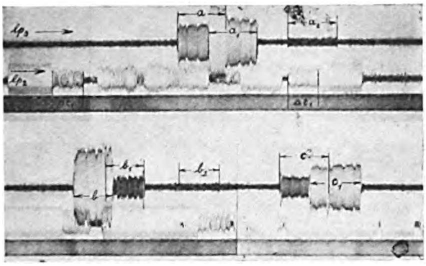

Fig. 11—Simultaneous records of Baires, LP3 (19,608 kc, or 15.3 m) and LP2 (14.5m) Oct. 27, 1927, 9.05 A.M., C.E.T. Indirect double signals. $\Delta t_1 = 0.056$ sec. $\Delta t_2 = 0.138$ sec.

There are next a few oscillograms of indirect double signals. Fig. 14, showing records of indirect double signals from station WTT ($f = 18{,}750$ kc, or $\lambda = 16$ m), was made on December 16, 1927, at 2:30 P.M. (C.E.T.), and gives a time difference of the double signal a_1 with respect to the main signal a of $\Delta t = 0.095$ second.

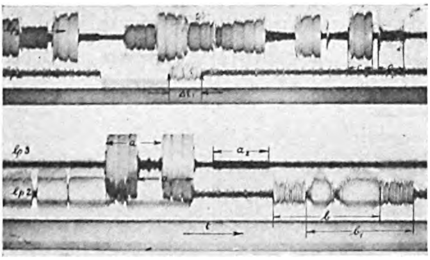

Fig. 12—Simultaneous records of Baires, LP3 (19,608 kc, or 15.3 m) and LP2 (20,690 kc, or 14.5 m) Oct. 27, 1927, 9.05 A.M., C.E.T. Indirect multiple signals. $\Delta t_1 = 0.056$ sec., $\Delta t_2 = 0.138$ sec. (On the right side perhaps Doppler effect.)

Fig. 15 contains indirect double signals of the station WTT, where in the lower picture b the double signals are somewhat weakened through the addition of a single reflecting system of wires. The time difference amounts in all cases to $\Delta t = 0.095$ sec. Similar oscillograms taken, however, simultaneously with and without screen arrangement, are shown in Fig. 16, station WLL, and Fig. 17, station 2XT, where the

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indirect double signals are eliminated sufficiently for practical purposes. Here two receivers were used with separate antennas, employing the same heterodyne circuit, in order to eliminate the disturbing influence of whistling. Fig. 18 gives another record from station 2XBC ($f=20{,}630$ kc, or $\lambda=14.54$ m) with indirect double signals which

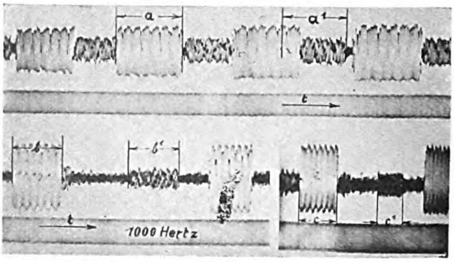

Fig. 13— Rocky point, WTT (18,750 kc, or 16 m), July 13, 1928, 2 A.M., C.E.T. Direct double signal, $\Delta t=0.138$ sec.

have the normal time difference of $\Delta t=0.095$ sec. with respect to the main signal.

III. China (Mukden)

So far only indirect multiple signals have been found from Mukden.

Fig. 19 contains two oscillograms from station XOM ($f=17{,}341$ kc, or $\lambda=17.3$ m) in Mukden, of November 3, 1927, 8:30 A.M. (C.E.T.). In the upper picture main and double signals, a and a_1, are equally large and result during overlapping in an approximately doubled

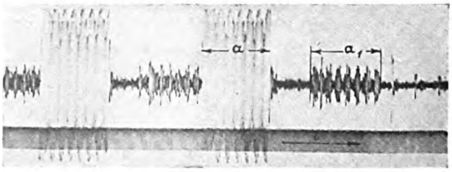

Fig. 14—Rocky Point, WTT (18,541 kc, or 16.18 m), Dec. 16, 1927, 2.30 P.M., C.E.T. Indirect double signal, $\Delta t=0.095$ sec.

amplitude. In the case of the lower signal b there is also to be distinguished a second indirect double signal b_2 with a time difference of $\Delta t_2=0.138$ sec. with respect to b_1. The indirect double signal b_1 has, with respect to the main signal b, a time difference of $\Delta t_1=0.086$ sec. The indirect double signals of Fig. 20 of the station XGA in Mukden ($f=15{,}000$ kc, or $\lambda=20$ m) are, in contrast to the previous picture,

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separated from the main signals, and show again a time difference of 0.086 sec.

Fig. 21, with a weak indirect double signal from station XGA, Mukden ($f=21{,}410$ kc, or $\lambda=15.01$ m), was taken at a time when we expected direct double signals. Further details are discussed in Sec. 4.

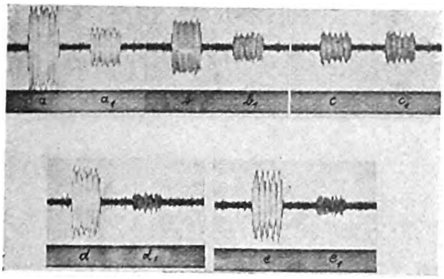

Fig. 15—Rocky Point, WTT (18,541 kc, or 16.18 m), Oct. 25, 1927, 2.35 P.M., C.E.T. Indirect double signals, $\Delta t=0.095$ sec. Records marked *a* are without reflector, *b* with reflector.

IV. Manila

Also on the line Manila-Berlin only indirect double signals could be traced in the records, and here again the amplitude of the double signal was usually greater than that of the main signals which had covered the shorter path. The time difference always amounted to 0.070 sec.

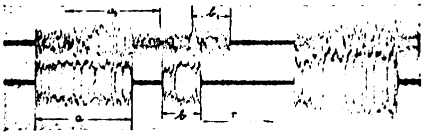

Fig. 16—Rocky Point, WLL, middle of January 1927. Indirect double signals with reflector, *b*; and without, *a*.

Fig. 22 is a record of indirect double signals from KZEN ($f=20{,}000$ kc, or $\lambda=15$ m) at Manila, in which the signals run completely together. A similar record of the 10,000-kc wave from KZET is shown in Fig. 23 for January 30, 1928, in which the much weaker main signal is considerably distorted, apparently because of rapid fading. The time difference is measured with considerable certainty as $\Delta t=0.070$ sec. Further records from Manila need not be given, since they offer nothing new.

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V. Java (Bandoeng)

The great circle Bandoeng-Berlin might be considered unfavorable for double signals, since Berlin in the northern hemisphere and its antipodal point in the southern hemisphere have about the shortest distance from the polar circles on the great circle. Hence only indirect double signals were noted, the photographic reproduction and analysis of which is very difficult in general, since the Java stations are excep-

Fig. 17—Rocky Point, New York, 2XT (18,750 kc, or 16 m), Dec. 30, 1926, 3.20 P.M., C.E.T. Indirect double signal ($\Delta t = 0.095$ sec.), with and without reflector.

tionally strongly modulated, so that the beginning and cessation of the signals cannot usually be exactly determined. Hence only a few records are of practical value.

Fig. 24 shows weak indirect double signals of the Java station in Bandoeng ANK ($f = 18,072$ kc, or $\lambda = 16.6$ m) with a time difference of $\Delta t = 0.062$ sec. Fig. 25 gives the same time difference, where the double signals stand out somewhat more clearly and appear in the further course of the record at times stronger than the main signals.

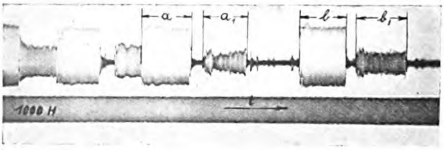

Fig. 18—Rocky Point, New York, 2XBC (20,630 kc, or 14.54 m), Jan. 31, 1928, 3.30 P.M., C.E.T. Indirect double signals, $\Delta t = 0.095$ sec.

VI. Siam (Bangkok)

After the establishment of a Telefunken short-wave transmitter in Bangkok early in 1928, rather strong double signals were sometimes observed, the amplitude of which in most cases was much larger than that of the main signals, which had covered a much shorter distance. Figs. 26 and 27 give examples of indirect double signals from the Siamese station HHP ($f = 17,751$ kc, or $\lambda = 16.9$ m) in which the time difference in both cases amounts to $\Delta t = 0.077$ sec. The station was working as a broadcasting station and had no directional antenna.

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VII. Japan

Also in the case of Japanese transmitters there were at first difficulties, as a result of poor modulation and lack of power, in observing the appearance of double signals. One record is given here, Fig. 28, from the Japanese station JAA, Iwaki ($f=12{,}050$ kc, or $\lambda=24.9$ m)

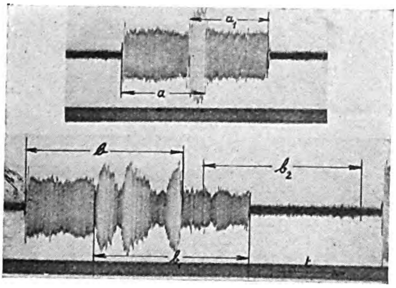

Fig. 19—Mukden, XOM, (17.3 m), Nov. 3, 1927, 8.30 A.M., C.E.T. Indirect double and multiple signals. $\Delta t_1=0.086$ sec., $\Delta t_2=0.138$ sec. (During the overlapping Doppler effect.)

which contains an indirect double signal with greater amplitude than the main signal and a time difference $\Delta t=0.076$ sec.

VIII. Nauen

Round-the-world double signals from our own Nauen transmitters were often observed and photographed. Here it was shown that the time difference, which corresponds to a circumference of the earth, remained constant within 1/10,000 sec. throughout the year. Variations at first (early in 1927) were due to incomplete facilities for

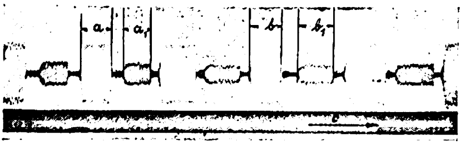

Fig. 20—Mukden, XGA (15,000 kc, or 20 m), Mar. 9, 1928, 8.00 A.M., C.E.T. Simple indirect double signals, $\Delta t=0.086$ sec.

measurement. The time difference varied between 0.138 and 0.1385 sec. A few oscillograms are presented.

Fig. 29 shows a complete round-the-world signal from the Nauen station AGA ($f=20{,}070$ kc, or $\lambda=14.95$ m) where in the case of the double signal, the irregular beginning of the main signal is very clearly shown.

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Fig. 30 shows a corresponding record from station AGC ($f=16,465$kc, or $\lambda=18.22$ m) with the same time difference. Fig. 31 may be of greater interest, as it shows a round-the-world signal from the experimental station AGK ($f=24,920$ kc, or $\lambda=12.04$ m), which station has also served for some time in commercial transoceanic communication. The time difference here also amounts to $\Delta t \doteq 0.138$ sec.

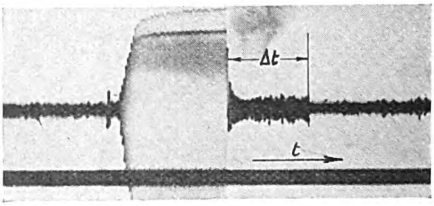

Fig. 21—Mukden, XGA (19,980 kc, or 15.01 m), July 16, 1928, 10.15 P.M., C.E.T. Indirect double signals, $\Delta t=0.0855$ sec.

In Fig. 32 another complete round-the-world signal from AGA is given, where the letter *v* at a sending speed of 150 w.p.m. appears as a double signal separated from the main signal.

IX. South Africa

There follows one more record, Fig. 33, of an indirect double signal from the Marconi station VNB ($f=18,750$ kc, or $\lambda=16$ m) in Cape Town, which we shall refer to in Sec. 4 in commenting upon the paper by Taylor and Young; the time difference is $\Delta t=0.072$ sec.

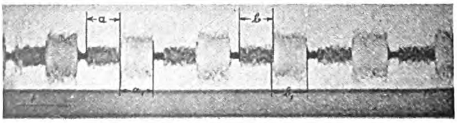

Fig. 22—Manila, KZEN (20,000 kc, or 15 m), Apr. 26, 1928, 8.40 A.M., C.E.T. Indirect double signals, $\Delta t=0.070$ sec.

4. Occurrence of Double Signals for the Various Directions, as Dependent upon the Time of Year and Day

Soon after the first observation of multiple signals from a North American station in the fall of 1926, the question arose concerning a certain regularity in their occurrence, especially because the double signals at certain times cause great disturbance in practical operation, when their amplitude exceeds about 30 per cent of the main signals. No conclusion could be drawn at that time, since there were but few

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short-wave transmitters available which operated regularly. Moreover, at that time commercial transoceanic communication with short waves was carried on only experimentally and to a very small extent. Hence, it was all the more important to determine in advance, with the appearance of new stations and new short-wave routes, the times for double signals.

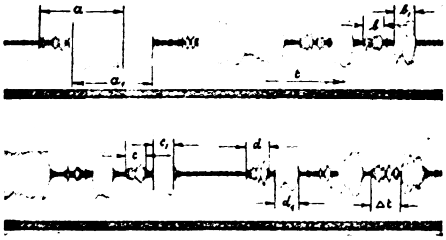

Fig. 23—Manila, KZET (10,000 kc, or 30 m), Jan. 30, 1928, 10.40 A.M., C.E.T. Indirect double signals which are stronger than the direct. $\Delta t = 0.070$ sec.

(*a*) *Twilight*: *Multiple signals*

A glance at the globe showed first of all, purely empirically, that some double signals were always present whenever the great circle connecting transmitter and receiver lay as close as possible to the twilight belt.

During the past year it has been discovered that the signals can pass around the earth several times only when the great circle lies in the twilight zone. The probability that encircling takes place more

Fig. 24—Java, (Bandoeng) ANK (18,072 kc, or 16.6 m), Feb. 14, 1928, 10.30 A.M., C.E.T. Indirect double signals, $\Delta t = 0.062$ sec.

than twice is greater the more closely the plane of the great circle approaches the twilight plane.

In order to determine whether the times of appearance of multiple signals coincide with the times of the passage of the great circles through the twilight zone, we must consider first somewhat more closely the relations between the various directions.

We consider first the three possible extremes:

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(1) The great circle, on which transmitter and receiver are situated, touches the polar circles:

In this case it is evident that the great circle on the longest, as well as on the shortest day, i.e. in summer and winter, at a definite time of

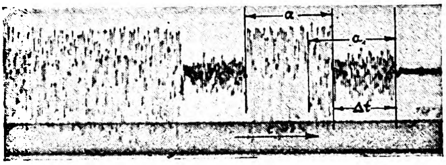

Fig. 25—Java (Bandoeng), ANK (18,072 kc, or 16.6 m), Aug. 31, 1928, 11.15 A.M., C.E.T. Indirect double signals, $\Delta t = 0.062$ sec.

day, depending on the location of the circle, coincides exactly with the twilight circle. Here it should be noted that the two times of day differ in summer and winter by twelve hours.

(2) The great circle passes through the poles:

Here the two circles coincide twice per day at the spring and fall equinoxes, at 6 A.M. and 6 P.M. sun time.

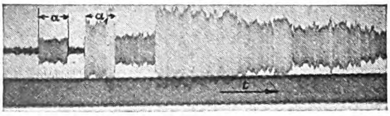

Fig. 26—Siam (Bangkok), HHP (17,741 kc, or 16.91 m), May 9, 1928, 11.10 A.M., C.E.T. Indirect double signals, $\Delta t = 0.077$ sec.

(3) The great circle passes along the equator:

In this case the circles can never coincide; the angle ϕ, which the twilight plane forms with the plane of the great circle, can only reach a minimum, once each on the longest and shortest day.

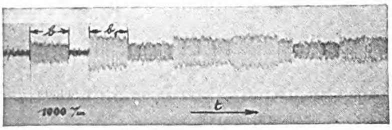

Fig. 27—Same as Fig. 26, only on Apr. 9, 1928, 9.05 A.M., C.E.T.

Hence, one can say that all of the great circles, which lie outside of the polar circles, are located nearest to the twilight zone on the longest and shortest day, *i.e.*, at a definite time of day once each in summer and

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winter. At this time the angle ϕ becomes a minimum. If the great circles touch the polar circles, the minimum angle ϕ becomes equal to zero at the same time. For all great circles which cut the polar circles the angle ϕ becomes equal to zero four times a year at a definite time of day, and these times approach the equinoxes as the great circles approach the poles. Let us now consider those great circles which are possible through Geltow (Berlin). (See also the map, Fig. 2).

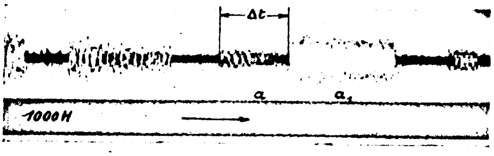

Fig. 28—Iwaki (Japan), JAA (12,050 kc, or 24.9 m), Dec. 7, 1927, 8.30 A.M., C.E.T. Indirect double signals, $\Delta t = 0.076$ sec.

ϕ_{min} can be here in the most unfavorable case = (90 deg. − 52.5 deg.) − 23.5 deg. (inclination of the earth's axis) = 14 deg. It has been found that multiple signals also occur when the twilight and great circles do not coincide exactly, owing perhaps to the finite width of the zone. As an upper limit, an angle of discrepancy of about 10 deg. has been found in practice. There is given below a tabular summary (Table I) in which for the main great circles, which correspond to the communication lines of the Transradio and a few other observed stations, the times at which angle ϕ reaches a minimum are recorded. At these times the great circles are located closest to the twilight zone. In addition the minimum value of ϕ is given.

TABLE I

Great Circle between Geltow and	Summer June 21 .E.T	Winter Dec. 12 C.E.T.	ϕ_{min} approx.
New York	2:30 A.M.	2:30 P.M.	−9.5 deg.
Rio de Janeiro	9:00 P.M.	9:00 A.M.	−4.5
Buenos Aires	9:00 P.M.	9:00 A.M.	−6
Mukden	9:00 P.M.	9:00 A.M.	−4.5
Osaka	8:30 P.M.	8:30 A.M.	−2
Manila	9:30 P.M.	9:30 A.M.	−10.5
Java	11:00 P.M.	11:00 A.M.	−14
San Francisco	4:00 A.M.	4:00 P.M.	+3
Bangkok	10:00 P.M.	10:00 A.M.	−13.5
Capetown	Spring 6:00 A.M. 6:00 P.M.	Fall 6:00 A.M. 6:00 P.M.	(about) 4 deg. from the pole

At the end of 1926 there were, as already mentioned, only a few short-wave transmitters available, so that the above assumption could not be tested until last year. For New York the time of appearance of double signals agreed with the assumed time very well. In the fall of 1926 and in the winter 1926–27 the times for the North American

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stations 2XT, 2XS and WIK were grouped around 2:30 P.M., C.E.T. The double signals first appeared weakly, 1/2 to 2 hours before their maximum, increased and reached their maximum at about 2:30 P.M., C.E.T., and then decreased for about the same length of time. A few weeks before and after the time of the shortest day their amplitude amounted at times to 100 per cent of the amplitude of the main signal.

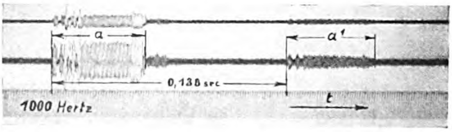

Fig. 29—Nauen, AGA (20,070 kc, or 14.95 m), Apr. 28, 1927, 10.04 P.M., C.E.T. Round-the-world signals, $\Delta t = 0.138$ sec., with a short range echo.

With the appearance of fading the main signal was at times weaker than the double signal. But it should be remarked that the fading was in general much stronger with the double signal, so that there were often difficulties in registering photographically on the oscillograph double signals on a film five meters long. Also the time 2:30 A.M., C.E.T., agrees very well for the occurrence of double signals from New York in the summer, as can be seen in the oscillogram, Fig. 13, from WTT. The direct double signal from WTT ($f = 18{,}750$ kc, or $\lambda = 16$ m) was registered at 2:00 A.M., C.E.T. While the time difference in winter corresponded to the indirect path, in summer the complete circle time of 0.138 sec. was registered throughout. The double signals

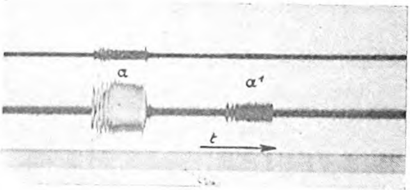

Fig. 30—Nauen, AGC (16,465 kc, or 18.22 m), Feb. 4, 1927, 10.40 A.M., C.E.T. Round-the-world signals, $\Delta t = 0.1385$ sec.

from New York appear, accordingly, to favor the indirect route in winter, the direct in summer.

After operations were begun with a new short-wave transmitter in Rio de Janeiro in the spring of 1927, it was very interesting to register the times for double signals and compare them with those calculated. Actually double signals occurred at about 8:00 P.M., C.E.T., reached a maximum at 9:30 P.M., and had disappeared about 10:30

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P.M. Unfortunately, the short-wave reception was often impaired at this time for weeks, as a result of magnetic disturbances (maximum of the 11-year sunspot period), so that the double signals likewise generally appeared weaker, and often ceased for a long time. Also a few weeks before the longest day, on June 21, and a few days later, the horizontal component of the earth's magnetic field showed marked

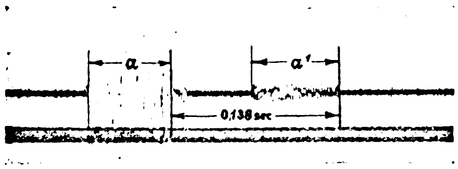

Fig. 31—Nauen, AGK (24,920 kc, or 12.04 m), Dec. 11, 1927, 11.20 A.M., C.E.T. Round-the-world signals, $\Delta t = 0.138$ sec.

disturbances, so that we waited in vain for the occurrence of multiple round-the-world signals, which we had long expected. Suddenly on June 25 multiple signals could be clearly heard, and two days later they were registered on the oscillograph.[5] We reproduce here a few extracts from the record made at that time, Figs. 4 and 5, where as many as five-fold signals from an emitted wave-train can be recognized. Rio was requested at that time to send short waves about 1/20 sec. in duration at intervals of a half second, in order to receive every

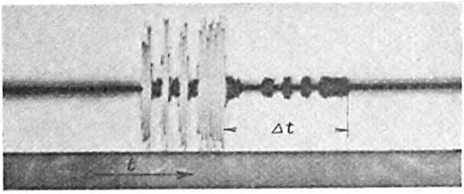

Fig. 32—Nauen, AGA, Sept. 19, 1928, 9.00 A.M., C.E.T. Direct round-the-world signal, letter *V* at sending speed of 150 w.p.m. $\Delta t = 0.1385$ sec., with short range echoes.

signal separately after encircling the earth. Unfortunately, Rio had misunderstood the communication and sent 2 dots per second with an interval of only 1/15 sec. In spite of this the signals can be analyzed clearly after passing three times around the earth. In Fig. 5 it is very interesting to note that the amplitudes of the double signals, a^1, a^2, a^3, decrease but little, quite in contrast for example, to the reception

[5] See also oscillograms in the paper by Quäck, *Elec. Nach. Tech.*, No. 4, p. 76, 1927.

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in a long reflecting telephone line where the amplitudes decrease regularly according to an exponential law. Even though momentary fading is present here, yet additional records taken successively so as

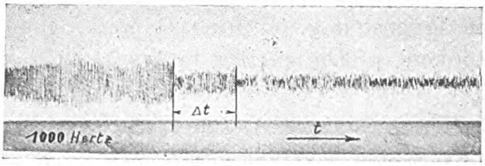

Fig. 33—Capetown, VNB (18,050 kc, or 16.62 m), May 14, 1927, 10.40 A.M., C.E.T. Indirect double signals, $\Delta t = 0.072$ sec.

to give a total length of 15 meters, are of about the same character. Some important conclusions can be drawn from this.

When in the course of 1927 and early in 1928 the short-wave stations LP2, LP3, and LP4 began operating in Baires, double and

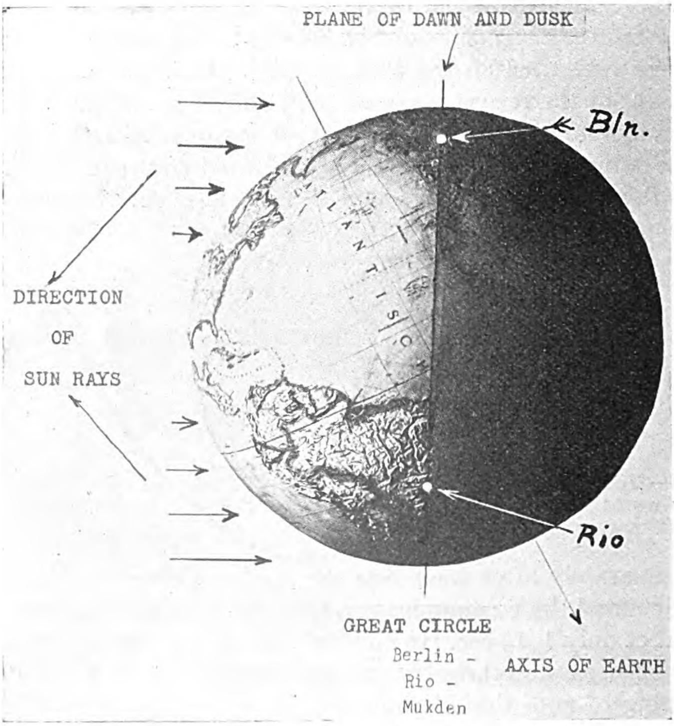

Fig. 34—The earth at the summer solstice, viewed from the side in a direction perpendicular to the sun's rays. The great circle Berlin-Rio-Mukden lies in the twilight zone at about 9.30 P.M., C.E.T.

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multiple signals could be observed at the same time as for Rio in the summer and winter, as an extract from the records, Figs. 6 to 12, shows. At the time of the shortest and of the longest day multiple signals were observed, in the summer only direct and in the winter indirect. This fact has but one meaning, and is extraordinarily interesting. When, late in 1927, in Mukden, which happens to lie approximately on the great circle Berlin-Rio, a Telefunken short-wave transmitter was installed, it was of interest to examine the direction and intensity of double signals coming from the east in addition to those from the west

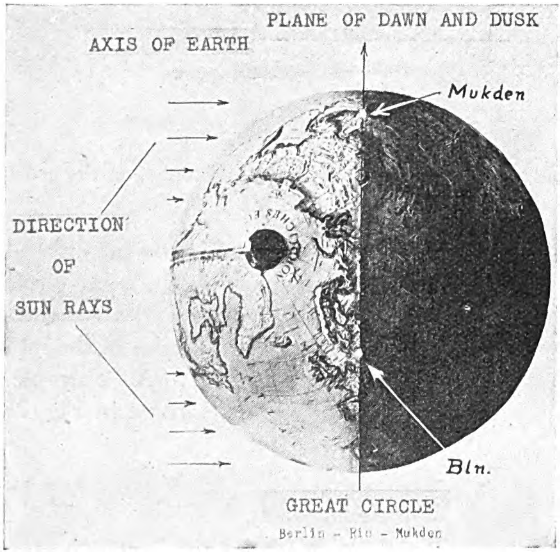

Fig. 35—The earth at the summer solstice, viewed obliquely from above. The great circle Berlin-Mukden lies in the twilight zone at about 9.30 P.M., C.E.T.

(Rio-Baires) which had been observed hitherto. The double signals expected in the winter at about 9:00 A.M., C.E.T., were registered exactly at that time; on the shortest day triple signals in the indirect direction were received. Examples are shown in Figs. 19 and 20, of November 3, 1927, and March 9, 1928. In the summer we expected, at about 9:00 P.M., C.E.T., direct double signals. Since Mukden was sending at this time with a lower frequency, and changing the frequency during operations gave rise to difficulties, the 20,000-kc frequency (15-m wave) could only be photographed once, on July 16, 1928, in which case strangely enough a weak indirect double signal (Fig. 21) appeared. This fact is extremely interesting and cannot be confirmed until next

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year, at which time Mukden will presumably have several transmitters. This one-day observation might be considered exceptional. No less scanty are the observations from Iwaki (Japan) in the east and San Francisco in the west. Here too the deviations of the great circles from the twilight zone are very small; ϕ_{min} amounts in the case of San

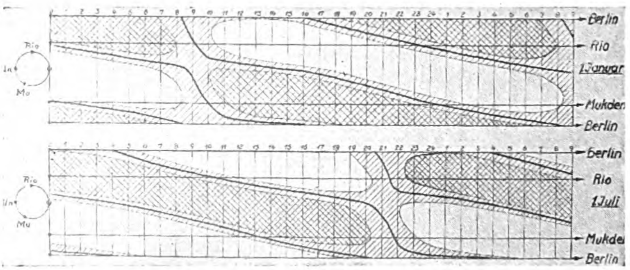

Fig. 36—Distribution of daylight, twilight, and night, on the great circle Berlin-Rio-Mukden-Berlin.

Francisco to only about +3 deg. in Iwaki (Osaka) only −2 deg., i.e., the great circle Berlin-San Francisco cuts the polar circles and coincides with the twilight zone shortly before and after the longest and shortest day, while the great circle Berlin-Osaka in the most favorable case comes very near to the twilight zone. An example of indirect double signals from Iwaki in the winter is given in Fig. 28 from JAA

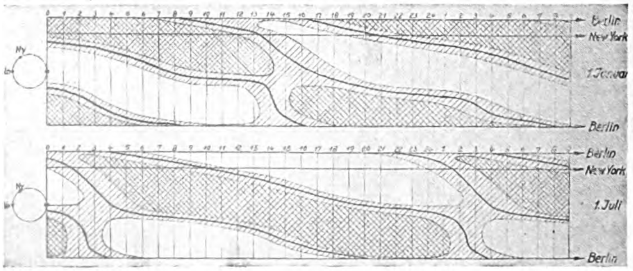

Fig. 37—Distribution of daylight, twilight, and night, on the great circle Berlin-New York-Berlin.

(December 7, 1927) about 8:30 A.M., C.E.T. The Japanese short-wave transmitters were, unfortunately, not working in the summer; however, late in September, 1928, direct double signals from the transmitter JAN in Nagoya (f = 11,720 kc, or λ = 25.6 m) were observed at about 8:30 P.M., C.E.T. The times recorded for San Francisco in Table I were also confirmed only a few times.

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We will now consider a few lines whose great circle through Berlin is far removed from the twilight zone, in which case ϕ_{min} is greater than 10 deg. Extended observations have been made from Bandoeng (Java), Manila, and Bangkok (Siam). Up to the present time only indirect double signals, that is, only simple echoes in the indirect direction have been observed, except from Manila, where also simple weak direct double signals were found in the three summer months. We might expect in summer weak direct double signals from the other stations mentioned also, if the stations operated with short waves at this time. For practical purposes the longer waves are in general better

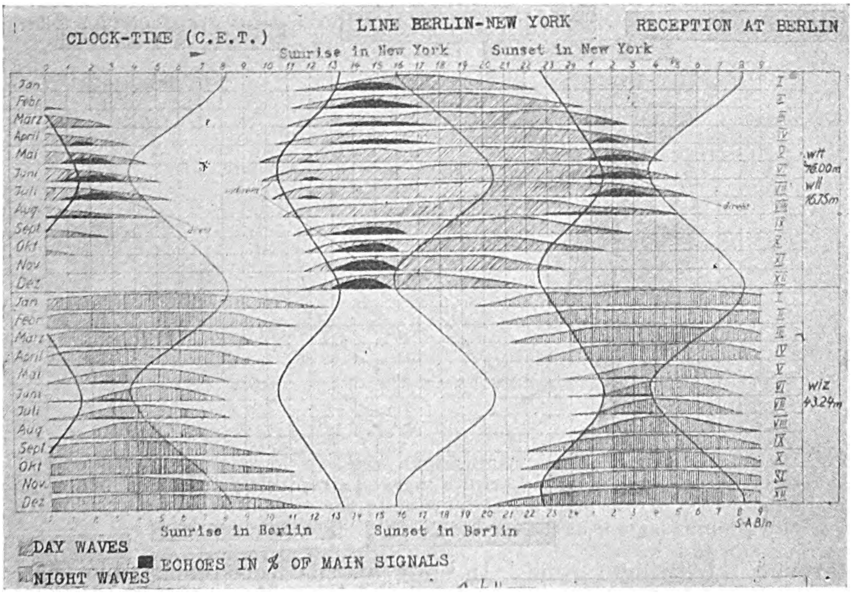

Fig. 38—Path from New York to Berlin. Diagram showing most favorable times for communication, and limits of audibility.

suited at this time. It must also be noted here that the strong indirect double signals in the winter fall more under the category treated of in Sec. 4b.

In the records of indirect double signals, Figs. 22 to 28 from Manila, Bangkok, and Iwaki, it is striking at times that the indirect double signals have a greater amplitude than the main signals. This very remarkable and regular occurrence can be reconciled, however, with the fact that the propagation conditions, at least for Berlin, are more favorable in the winter in the indirect, in the summer in the direct sense. An explanation of this cannot be given as yet.

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The line Capetown-Berlin presents a special case, where the great circle cuts the polar circles and passes only about 4 deg. from the pole. Here the great circle passes through the twilight zone twice a year, in spring and fall, morning and evening. Extended observations actually showed that at about 6:00 A.M. and 6:00 P.M., C.E.T., at the equinoxes strong double signals from the transmitter VNB in Capetown were present. Their direction, as a result of high speed of transmission, has, as yet, been observed only once as indirect (see for example, Fig. 33), in which, as a result of poor conditions, it happens that the area of disturbance was very large. We can extend the observations of Taylor here to the extent that, even across the poles, double signals are possible, if at the time of equinoxes the great circles pass

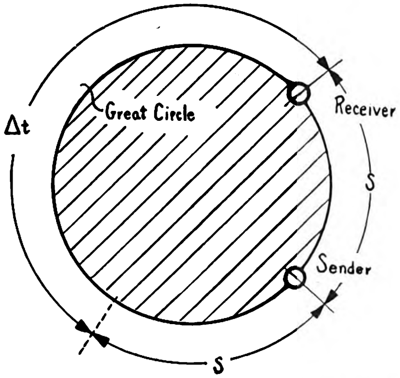

Fig. 39—Great circle of the earth, with relative locations of sending and receiving stations. Curved arrows marked *s* represent distances, while the long curved arrow marked Δt represents the time interval between direct and indirect signals as received.

through the twilight zone. In spite of this it is remarkable that, according to Taylor, as yet no double signals from Baires have been observed in New York, although the angle ϕ_{min} during the whole year is always smaller than 10 deg. Perhaps here the magnetic poles still have some influence. If one were to consider still other observations, which we have, though incomplete, from Baires, Rio, New York, and Mukden, one could almost come to the conclusion that the phenomena mentioned above are peculiar to certain parts of the earth. Some time will probably pass before exact continuous observations are obtained from all parts of the world.

Interesting results were obtained from the observation of complete round-the-world signals from our Nauen transmitters, which, it is true, were not always carried out quite regularly. Since the short-distance Nauen-Geltow (30 km) can be disregarded so far as direction is con-

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cerned, double signals are expected in every season at sunrise and sunset. When the stations were working with non-directive antennas, this could also be confirmed. For some time directional antennas have been used almost exclusively so that the times for double signals are somewhat more limited and hence fall in the periods where the directions used coincide with the twilight zone. Examples of double signals of the Nauen transmitters are given in Figs. 29 to 32.

To furnish a clearer picture of the passing of a great circle through the twilight zone, the globe with a corresponding position of the great circle Berlin-Rio-Mukden was photographed and is represented in Figs. 34 and 35.

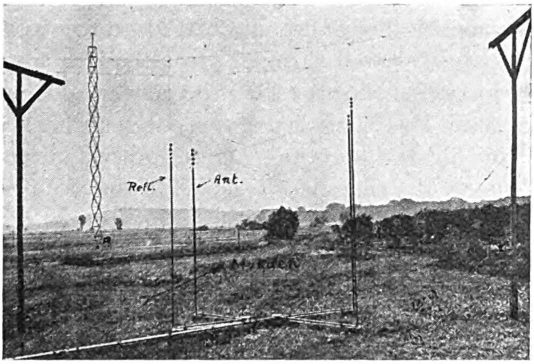

Fig. 40—Antenna systems at Geltow for directional reception from Mukden (14 and 19 meters), with electricall yexcited reflecting wires* for production of cardioid characteristic.

The pictures represent the longest day (summer), when the North Pole is farthest from the twilight zone. They consist of separate views, from the side (Fig. 34) and obliquely from above (Fig. 35), in order to permit the three places Berlin, Rio, and Mukden to be recognized. The time is about 9:30 P.M., C.E.T. In Rio and Berlin the sun is just setting, in Mukden rising. At this time the signals from Rio appear very strong. In winter on the shortest day the situation is just reversed; the sun's rays then pass parallel from right to left on the picture and light and dark are reversed. The time is then changed 12 hours to 9:30 A.M., C.E.T., when the indirect multiple signals from Rio and Mukden appear very strong.

(*b*) *Southern Hemisphere in Summer at Night: Indirect Double Signals*

It was soon found that double signals sometimes appeared outside of the twilight zone; in this case it was always a matter of indirect

* "Mit gespeisten Reflectordrähten."

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double signals, which cover the usually longer reverse path, while hitherto direct, as well as indirect, multiple signals had been registered only in the twilight zone. These double signals usually take the reverse route in the summer at night; at that time the complete direct path lies entirely in the light and 20,000 km of the reverse route in the dark. Let us consider first 25,000 to 13,636-kc frequencies (12 to 22 m), which we designate as *day* and *transition* frequencies because the higher frequencies of about 25,000 to 16,667 kc (12 to 18 m) are better suited for day communication, where the complete route lies in the light while the somewhat lower frequencies, 16,667 to 13,636 kc (18 to 22 m), are suited to the transition period between day and night, where part of the path is in the light, the rest in the dark. The term "day wave" is, however, not quite appropriate, for in summer these waves pass just as well through areas that are absolutely dark. It must not be forgotten that for the propagation of short waves, and for the height of the Heaviside layer, not only sunrise and sunset and the division of the day into daytime, night, and twilight are influential, but also particularly the change of season. In the tropics the so-called day waves are just as well suited for communication at night over very great distances.

If the route in question crosses northern or southern areas, then the change between summer and winter becomes much more conspicuous. In winter, so far as short waves are concerned, transmission is possible only during the day. It is easy to see, however, that then the transmission can also take place on the reverse night path, if the summer season prevails there. That is, however, naturally always the case for a few hours; when, for example, it is winter in the northern hemisphere, the southern hemisphere is in the summer season, so that for most great circles which pass through Berlin such cases are possible in the winter, when the short waves can pass along the usually much longer reverse route because of the low Heaviside layer.

A few examples are now given for the indirect double signals discussed above. From South America (Rio and Baires) these double signals appear in the winter months between 11:00 A.M. and 7:00 P.M., C.E.T. Their amplitude is, disregarding magnetic and other influences, extraordinarily large, and often reaches values of 100 per cent of the main signals (compare Figs. 8 and 9). Fading is usually stronger and lasts longer than with the direct signal. In this kind of indirect double signal a special regularity or order is seldom recorded, in contrast to the twilight double signals discussed in the preceding section. Their amplitude, for example, can be just as large at the time of equinox, as at the time of the shortest day. In the four summer months, when the

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southern hemisphere has winter, these indirect double signals cease.

The indirect double signals from Bangkok, Bandoeng, Manila, and Mukden are not so clear but can, nevertheless, be distinguished; these also appear, strangely enough, on into the summer, when a considerable part of the reverse path traverses regions where night and winter prevail. Still, with Siam, Java, and Manila, the explanation may be found in that these transmitters are located near the equator, where in the summer and winter the Heaviside layer is very low. From New York only twilight double signals were observed, in general, except recently, when after an increase of power of the transmitters in New York, exceptional cases, still unexplained, also occurred.

It should be noted also that from the European short-wave transmitters in London, Paris, Kootwyk, Lisbon, and Madrid, except at twilight during the winter months, indirect double signals were found. Likewise the signals sent out from Nauen can pass around the earth in the winter months in the forenoon, when the northern hemisphere has day and the southern has summer and night.

(*c*) *Reverse Path Better Suited for Certain Waves: Indirect Double Signals*

It can, of course, happen that the return path is better suited for the wave than the direct path. If the distance from transmitter to receiver ranges from 5000–8000 km, a night wave can still be audible on the direct light path, while the amplitude received over the reverse route, located mostly in darkness, is just as great or greater. Such a case is presented, for example, in the reception of the 10,000-kc frequency (30-m wave) from Manila (Fig. 23). Here strong indirect double signals are observed in the winter months between 9:00 A.M. and 12:00 P.M., C.E.T., while the direct reception is very weak. This time may fall in the twilight period, in which case the situation is easily explainable for the longer waves. Characteristic for this type of indirect double signals is the fact that they almost always occur at times when the reception on the direct path is just coming through or just ceasing. Similar results were obtained in the observations of longer waves from Bangkok and Mukden. The circumstances are similar for Australia (Sydney), which is located only about 4000 km from the antipodal point of Geltow, where the reception of somewhat lower frequencies of about 12,000 kc (25 m) in the transition periods is possible from both directions, but very weak.

In general it may be noted here that the amplitudes and field strengths of direct and indirect signals are much smaller than with the short waves in the twilight period. The amplitude ratio is roughly of the order of 1:10.

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(d) Strengthening of Indirect Double Signals by Reverse Radiation from the Transmitter.

In the first place, it must appear paradoxical when from those transmitters with antennas that are sharply directive toward the receiving station, there appear at the times given above *indirect* double signals. Usually the directional characteristics of these transmitters are determined either by calculation or else experimentally by means of small auxiliary transmitters on the ground. In practical communication over great distances, however, radiation appears always to be perceptible even in the reverse direction. The screening at the rear of these beam transmitters was not yet complete. In the meanwhile technical advances have been made, which allow complete elimination of reverse signals. An interesting observation was made when a new transmitter in Baires was put into operation. At a time when weak indirect double signals were to be expected *indirect* double signals, stronger than the direct ones, were observed. While two other Baires transmitters showed at the same time and at nearly the same wavelength weak indirect double signals, the amplitude of the indirect signal from the new transmitter was very materially stronger than that of the direct signal, which had only a third as much path to cover. After a few days the attention of Baires was called to this fact and inquiry was made there as to whether the direct radiation from their antenna was not accidentally screened off. The antenna of the new transmitter was then changed, whereupon normal conditions actually appeared. One sees from this that the indirect double signals can be strengthened by radiation artificially directed backwards.

(e) Location of Great Circles and Division of Day, Twilight, and Night

Since the times of sunrise and sunset, and the division of day, twilight, and night, along the great circle are of great importance in transmission, graphical charts were made for various great circles of interest, of which two are presented in Figs. 36 and 37. In Fig. 36 the great circle Berlin-Rio-Mukden is considered, in Fig. 37 the circle Berlin-New York.

The circle has been unrolled and the abscissas represent "clock-time," while the ordinates represent distances. The heavy lines represent the border between sunrise and sunset. Night areas are indicated by heavy shading, twilight areas by light shading, while day areas are left clear. The light lines mark the border between twilight and night or day. Since twilight is not sharply defined, those points were taken as the limits in which the sun, as observed at a height of 200 km, would be rising or setting. One obtains thus a clear picture at least as

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to points with the same illumination. The values were obtained from astronomical tables. Each diagram refers to a particular day, and approximately the most extreme cases were chosen, namely, the longest and shortest days. It is seen from Fig. 36, for example, that the entire great circle Berlin-Rio-Mukden lies in the twilight zone on January 1 about 9:00 A.M., C.E.T., on July 1 about 9:00 P.M., C.E.T.; at these times the multiple signals actually occurred. The same is true of New York in Fig. 37 on January 1 about 2:30 P.M., C.E.T., and July 1 about 2:30 A.M., C.E.T. These charts were made for every month, so that from this the amount of light along the path can be calculated at all times of day and year. The amplitudes of reception are not included in these charts to avoid confusion.

(f) Graphical Representation of Amplitudes of Reception and Their Dependence upon the Time of Day and Year.

The advantages of certain waves for the various directions is brought out clearly in graphical charts of the reception which we have prepared for most of the lines. These charts of the most favorable transmission times and limits of audibility summarize our investigations extending over three years, and contain valuable material for research in propagation phenomena.[6] As an example we present in somewhat different form in Fig. 38 the strength of received signals at day frequencies (18,750 kc or 16 m) and at night frequencies(9,090 kc or 43 m) from New York. The wavy lines running in the direction of the ordinate represent the time of sunrise and sunset at the transmitting and receiving stations as calculated from astronomical tables. The *monthly averages* of the signal strength are shown as related to the time of day. The audibilities are approximately proportional to field strengths and were measured uniformly by means of a Barkhausen audibility meter. The measurements took place at the beginning of the Geltow-Berlin land-line, where approximately constant conditions always prevail. Naturally only relative values are obtained, which suffice, nevertheless, for a judgment concerning transmission possibilities. The amplitudes are plotted approximately to scale, that is, the interval from month to month corresponds to an audibility of 5000 Wien* (units of audibility). In order to avoid the use of two small

[6] The material was published in *Elec. Nach. Tech.*

* Wien units according to Barkhausen-

Wien	Audibility
1 - 2	R 1
4 - 8	R 2
16 - 32	R 3
64 - 125	R 4
250 - 500	R 5
1000 - 2000	R 6

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units, transmission values of over 4000 Wien are not indicated, since this received energy is sufficient in every case to insure good communication at high speed, so long as the transmitter is in good condition.

The double signals are indicated in black and are given in percentages of the amplitude of the main signal; the direction of the path, direct or indirect, is also indicated. Fading is not represented in this chart. Likewise the path between transmitter and receiver is not indicated here. One sees from Fig. 38 that the twilight double signals from New York are strongest when times of sunrise and sunset at the places of sending and receiving approximate each other most closely. Moreover, the possibility of good transmission is particularly pronounced in summer, while both stations are in darkness, corresponding to the low height of the ionized layer.

5. Range of Wavelengths in Which the Various Double Signals Appear.

For the sake of making the survey clearer, those frequency ranges will now be summarized in which the double and multiple signals mentioned above appear. For the twilight double signals we can select a range of 25,000 to 13,636 kc (12 to 22 m) perhaps even up to 12,000 kc (25 m). The most favorable frequencies are located apparently between 20,000 and 18,750 kc (15 and 16 m), where thus far as many as five-fold signals from a train of waves have been observed.

If one extends the meaning of "twilight double signals" somewhat farther and includes the indirect double signals noted under Sec. 4c, then the frequency range can be extended to the night frequencies also, up to about 6,667 kc (45 m). Thus in the twilight zone, direct and indirect double signals between 25,000 and 6,667 kc (12 and 45 m) were observed. The indirect double signals noted under 4b, seem, on the other hand, to be limited to a smaller range of frequency, approximately 21,430 to 13,636 (14 and 22 m). Here too a maximum between 20,000 and 17,647 kc (15 and 17 m) is found, yet a similar signal is found for certain directions at 14,286 kc. If the sending place is very near the antipodal point of the receiver, then the so-called transition waves, i.e., waves between 13,636 and 10,345 kc (22 and 29 m) are well suited for the indirect double signals which appear (e.g., Australia: 12,000 and 10,345 kc, or 25 and 29 m).

For the indirect night double signals at greater distances, where transmitter and receiver are separated about one fourth of the earth's circumference and less, a range of 10,000 to 8,820 kc (30 to 34 m) is, in general, suitable.

6. Measurement and Calculation of the Time-Differences

In contrast to the results of Taylor and Young, which show varia-

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tions in the time differences, with the same transmitter, amounting sometimes to over 10 per cent, our measurements of time differences generally lie within the limits of errors of observation and the precision of the apparatus. The differences usually amount to 1 per cent or less. Occasionally there were greater variations which could, however, always be traced back to a change of frequency in the 1000-cycle generator.

One can probably assume as certain that the path of the multiple signals, even when the plane of polarization is rotated, is always along the great circle connecting transmitter and receiver.

From the time difference of multiple signals with respect to the direct main signal and to each other, it is easy to determine in which direction the waves have passed, whether directly or indirectly. Long continued observations of the double signals of our Nauen transmitter in Geltow showed a mean time difference of 0.138 sec.; when the signals were sent with great exactness even 0.1385 sec. could be measured. If one neglects the change in propagation velocity in the atmosphere, then this time difference corresponds to about 41,400 km, i.e., a route 3.5 per cent longer than the earth's circumference.

We next computed the time differences between direct and indirect signals for the various transmission paths, on the assumption that all distances were to be multiplied by 1.035.

The distances were calculated by the Geodetic Institute in Potsdam (Prof. Dr. Mahnkopf) within about 10 km, and kindly placed at our disposal. If *s* designates the distance along the shortest arc in km, then the time difference between direct and indirect signals is given by the following formula:

$$\Delta t_{\text{calc}} = \frac{(40020 - 2s) \cdot 1.035}{300{,}000}$$

A great circle showing relative positions of transmitter and receiver is represented in Fig. 39. In Table II the calculated and measured time differences are given, as well as the actual distances within 10 km.

TABLE II

Path Geltow to	kc	or m	Distance (km)	Δt calculated (sec.)	Δt measured (sec.)
Nauen	27,270–11,538	11–26	40,020	Assumed dist. s+3.5 per cent	0.138–0.1385
New York	21,430–13,043	14–23	6,390	0.0942	0.094–0.095
Rio	20,550–19,231	14.6–15.6	9,940	0.0695	0.069–0.070
Baires	20,980– 8,570	14.3–35	11,850	0.0562	0.056
Java (Bandoeng)	18,750	16	10,930	0.0627	0.062
China (Mukden)	20,000–15,000	15–20	7,630	0.0856	0.086
Japan (Iwaki)	15,000	20	8,870	0.0769	0.076
Manila	10,000–20,000	30–15	9,900	0.0698	0.070
Siam (Bangkok)	21,430	14–17	8,650	0.0783	0.078
Capetown	18,250–17,647	16	9,590	0.0721	0.072

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It is striking how well the calculated values agree with those photographed on the oscillograms.

If, as stated recently by H. T. Friis,[7] it is true that the horizontal angle of reception varies as much as 30 deg. from the geographic direction, then it may be either that the corresponding time differences are less than the errors of observation, i.e., less than 10^{-3} sec. (perhaps as a result of local conditions), or that such great variations occur but seldon, so that they were not measured by us in spite of regular photographing.

7. Effect of Magnetic Disturbances

Short-wave reception suffers in general not only from fading and from the marked influence of the illumination of the path, but particularly from magnetic disturbances. The most important causes are probably found in the appearance of sun spots, since the periodicity of short waves and of magnetic disturbances is clearly related to the rotation of the sun. The disturbances seem to be very effective for all directions approximately perpendicular to the lines of the earth's field.

With the appearance of magnetic disturbances, the multiple signals disappear in every case, while the amplitude of double signals is more or less strongly affected according to the direction. Thus, for example, even with weak magnetic disturbances, no double signals are observed from New York, whereas from South America they still occur, since the circle Berlin-Rio-Baires makes a very small angle with the magnetic lines, in contrast to New York where the great circle cuts the lines almost perpendicularly. With moderate and stronger disturbances the signals from South America also disappear.

8. Elimination of Double Signals

It is already evident that the elimination of *direct* double signals, or their exclusion from the receiving antenna, is probably not po sible, in contrast to the *indirect* double signals, which strike the rec^iving antenna from the opposite direction. An elimination of the in^irect double signals is achieved by setting up an aperiodic grounded screen, which can consist of several single wires. Here it is to be noted that in the larger receiving stations under certain circumstances several antennas are set up close together, in which case the incident waves of the double signals are scattered diffusely, so that in spite of the best side screening, effects are still noticeable. As Figs. 16 and 17 show, the screening with such simple apparatus as that mentioned above is completely successful. A further step is reached through the use of

[7] H. T. Friis, Proc. I. R. E., **16**, 658; May, 1928.

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tuned reflecting wires back of the receiving antenna, in which case by much simpler means the same, or better, elimination is obtained. Recently special reflector apparatus has been introduced, which results in an exact cardioid characteristic; except for the Beverage antenna, this probably offers the simplest means at present for keeping indirect double signals away from the receiver.[8]

Fig. 15 shows comparative records, obtained with and without simple reflector wires, where the amplitude of the double signals, as a result of poor tuning, was reduced only about one third.

[8] Fig. 40 shows two such directional antennas built at Geltow by Telefunken for the reception from Mukden.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

SHORT RANGE ECHOES WITH SHORT WAVES*

By

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(Transradio, Berlin, Germany)

***Summary**—The authors have made an examination of short range echoes by means of special signals transmitted from the high-frequency stations at Nauen (21,430 to 11,538 kc, or 14 to 26 m) and recorded photographically at Geltow. The echoes are sometimes blurred and sometimes repeated at fairly regular intervals, indicating multiple reflection between the ground and a reflecting layer. In one case the height of this layer is estimated to be about 1,500 m. The character of the echoes may change radically within a few seconds. General agreement is found between the authors' observations and those of Taylor and Young and also those of Hoag and Andrew. The bearing of the echoes on radio communication is discussed and a number of oscillograms are reproduced.*

In the summary of the paper by one of the present authors on "Double and Multiple Signals on Short Waves,"[1] par. 16, reference was made to short range echoes (Nahechos). Concerning these we wish to make the following report.

On the photographic records of round-the-world signals from the Nauen transmitter made at the receiving station in Geltow it frequently happened that the signals were accompanied by small, ob-

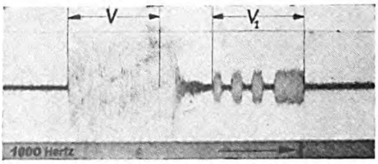

Fig. 1—Records made at Geltow of V signals from the Telefunken Transmitter AGA (20,090 kc) in Nauen, Nov. 14, 1928, 11.10 C.E.T. The direct signal V is blurred by short range echoes, while the round-the-world signal V_1 is reproduced exactly.

scure appendages, sometimes at the beginning and sometimes at the end of the signals. Since these appendages could also be recognized with certainty in the double signals as well, the phenomenon was considered an indication of faulty keying. Later it was recognized that these appearances were partly attributable to short range echoes. This term was chosen because these wave groups occur in shorter time

* Dewey decimal classification: R113. Original manuscript received by the Institute, December 27, 1928. Translation received, January 11, 1929. Supplement to "Double and Multiple Signals on Short Waves," Proc. I. R. E., 17, 791; May, 1929.

[1] *Elek. Nach. Tech.*, 4, p. 74, 1927.

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intervals than is the case with the direct or indirect double signals, which pass completely around the earth in one direction or the other.

The first positive proof of short-range echoes from the Nauen transmitter was furnished when the signals received in Geltow from this station were completely blurred and irregular, while at the same time New York and Baires were copying the same transmitter at

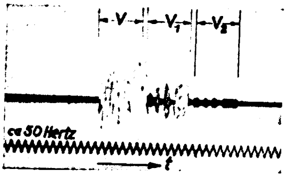

Fig. 2—V signals from AGA in Nauen (20,090 kc, or 14.93 m) received in Geltow Nov. 17, 1928, 10:40 A.M., C.E.T., showing double round-the-world signals V_1 and V_2. Owing to short range echoes, V_1 is merged into V.

Nauen at the rate of over one hundred words per minute. In order to investigate this point a series of widely spaced V signals were sent from Nauen at a rate of one hundred and twenty words per minute. At Geltow there were recorded, in addition to the greatly blurred, direct V signals, one or two round-the-world echoes which were absolutely clear and above criticism. Some examples of the oscillograms are reproduced in Figs. 1 to 3. In Figs. 1 and 2, V, the short range

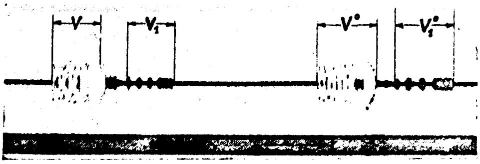

Fig. 3—Record of Nauen transmitter AGA1 made at Geltow (20,190 kc, or 14.86 m), with round-the-world signals V_1 and V°_1, in which the short range echoes appear separated.

echoes merge with the main signal, forming a complete blur. As a result of the short range echoes the total duration of the main signal V is about 0.04 sec. greater than that of the double signal V_1. In Fig. 2 the round-the-world signals can be clearly recognized. The distance from V_1 to the main signal V should be the same as that of the multiple signal V_2 from V_1; nevertheless it will be noted that the beginning of V_1 overlaps the short range echoes somewhat. It is clear from the photographs that these short range echoes, judging from the time intervals, cannot have passed around the earth. In order to investigate

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this matter somewhat more closely, signals were sent in intervals of ½ sec. from the Nauen transmitter AGA1 (20,440 kc, or 14.68 m), AGA (20,090 kc, or 14.93 m), and AGB (11,380 kc, or 26.36 m) in succession. For this purpose a tape was prepared containing a repetition of the letter *e* separated by ten to twenty spaces. This was sent through the machine transmitter at a rate of one hundred to one hun-

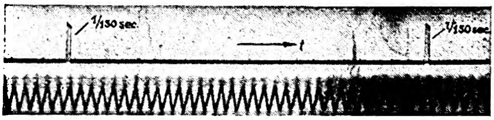

Fig. 4—Control record of AGA1 in Nauen, made in the transmitting room. Short dots of 1/150 sec. duration at intervals of about 1/2 sec. The keying is absolutely exact.

dred fifty words per minute. Special control registrations were also made in the same building in which the transmitter was located, in order to make certain that the transmitter actually was emitting short, exactly defined signals. One of these records is shown in Fig. 4.

The first of these exposures recorded at Geltow from AGA is shown in Fig. 5, in which, in addition to the direct signal *a* and the weak

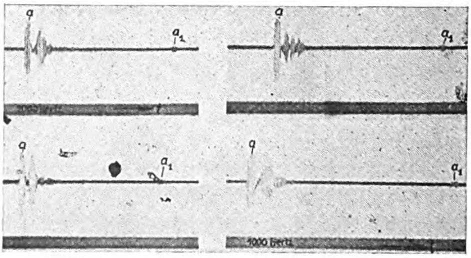

Fig. 5—Short range echo records from AGA in Nauen, Nov. 14, 1928, 12:30 P.M., C.E.T. The short range echoes, which follow the main signal *a* have quite varying character on the four records, all of which were made within 3 secs. a_1 is a round-the-world signal.

round-the-world signal a_1, between one and three short range echoes are present. These short range echoes present quite different appearances in the four exposures, all of which were made within three seconds. Some of them appear merged together while others are separated. By taking averages of time intervals from the entire film, which was five meters long, it was possible to distinguish three-time intervals of short range echoes with respect to the main signal:

$$\Delta t_1 = 0.0095 \text{ sec.}$$

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$$\Delta t_2 = 0.020 \text{ sec.}$$
$$\Delta t_3 = 0.0295 \text{ "}$$

As a consequence of the continually varying appearance of the echoes and of the fact that the time intervals in contrast to the round-the-world signals are very variable, it is possible to give only average values, but these averages are repeated with a fair degree of regularity. Fig. 6 shows particularly well how irregularly the reflections often occur. In this figure a round-the-world signal of great amplitude is also present. It is very helpful to have a round-the-world signal on the same record, since by its use the length of the emitted signal can be determined, as cannot always be done from the direct signal itself or from the short range echoes on account of blurring. The short range echoes would probably overlap even if much shorter dot-signals were emitted, since the reflections take place not only from

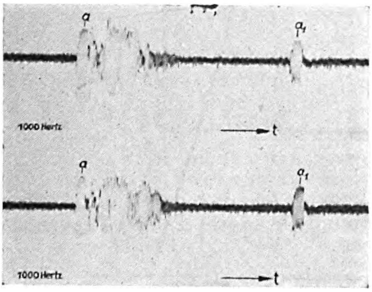

Fig. 6—Short range echoes received on non-directive antenna at Geltow, Nov. 15, 1928, 10:40 A.M., C.E.T., from AGA 1 (20,190 kc, or 14.86 m) in Nauen. Between the main signal a and the round-the-world signal a_1, which is of very great amplitude, the short range echoes merge together.

varying distances by discontinuous jumps, but also somewhat after the manner of peals of thunder. Nevertheless, there are also times when the short range echoes appear separated even when short signals of about 1/200 sec. duration are used. We have been particularly successful in making such records as these by using a directive antenna at the receiving station. An example of this is shown in Fig. 7, all three records having been taken from a single oscillograph film. The signal was received on a multiple antenna that was sharply directive toward the southwest. Nevertheless the amplitude of the direct signal from Nauen, which is located at a distance of about 30 km almost directly north from Geltow, is reduced to about a fifth part. On the other hand, the first short range echo appears particularly strong. On the whole, in addition to the round-the-world signal, seven short-wave

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echoes can be clearly recognized; the values of the time intervals were derived from many records, and the mean values are given in the following table, together with the corresponding distances in kilometers.

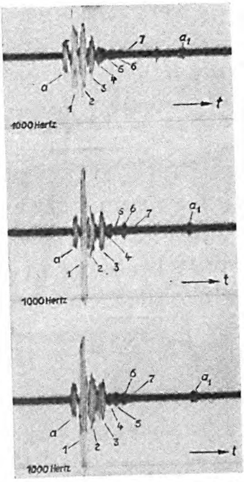

Fig. 7—Seven-fold short range echoes from AGA (20,090 kc, or 14.93 m), recorded with directive antenna in Geltow. In addition to the main signal a, the amplitude of which is reduced by the directive action of the receiving antenna, and the round-the-world signal a_1, seven short range echoes can be clearly distinguished.

As indicated in the last column of Table I, it was found almost always in the reception of AGA and AGA1 that the time intervals for a group of short range echoes were integral multiples of the time interval of the first echo. These regular time intervals, in which the short

TABLE I
TIME INTERVALS AND REFLECTION DISTANCES FOR SHORT RANGE ECHOES

Time Intervals (means)		Limits Seconds	Approximate Distance km	Approximate Multiples
Symbol	Seconds			
Δt_1	0.0105	0.085–0.012	3150	s_1
Δt_2	0.021	0.016–0.025	6300	$2s_1$
Δt_3	0.031	0.030–0.032	9300	$3s_1$
Δt_4	0.042	0.040–0.044	12600	$4s_1$
Δt_5	0.053	0.050–0.057	15900	$5s_1$
Δt_6	0.065	0.063–0.067	19500	$6s_1$
Δt_7	0.077	0.074–0.079	23100	$7s_1$

range echoes appear, lead to the conclusion that the waves are reflected several times between two layers. One of the reflecting surfaces must be the earth itself, while the other is a layer about 1,500 km in height. Short range echoes have also been observed by Taylor and Young, and they were investigated further somewhat later by Hoag

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and Andrew.[2] The time intervals given by them are of the same order of magnitude as our own and are in part in agreement therewith.

From the Nauen transmitter AGB (11,381 kc, or 26.36 m) short range echoes were also received at Geltow as indicated in Fig. 8. Hoag and Andrew have also found in Chicago evidence of short range

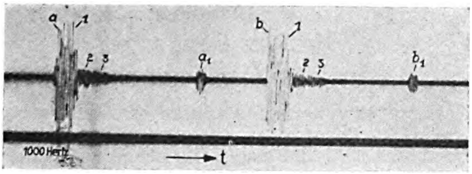

Fig. 8—Short range echoes recorded in Geltow from Transradio Transmitter AGB (11,381 kc, or 26.36 m), Nov. 16, 1928, 10:00 A.M., C.E.T. Between the main signal a and the round-the-world signal a_1 three short range echoes 1, 2, and 3 can be recognized.

echoes from this same transmitter. Their observed time intervals correspond throughout to those which we have measured, only with this limitation— that the time interval of 0.083 sec. given by Hoag and Andrew seems to correspond to a normal, indirect double signal. A comparison of the measured values is shown in the following table.

TABLE II

Time Intervals in Seconds

AGB Received in Geltow	AGB Received in Chicago
0.006 sec.	0.005 sec.
0.015 "	0.006 "
0.028 "	0.016 "
0.055 "	0.083 "

An examination of several hundred oscillograms has shown that occasionally short range echoes occur in the reception of signals from distant stations. Fortunately, in such cases their amplitude is usually so small that distant communication, in contrast to communication over short distances, hardly suffers as a consequence. On the other hand, direct double signals can be the occasion for disturbances over great distances.

It should be remarked that it is extraordinarily difficult in the reception from distant transmitters to separate the short range echoes from irregularities in transmitting. In most cases an exact discrimination is possible only through the study of oscillograms made simultaneously at both transmitting and receiving stations. For this reason a considerable time may still elapse before the phenomena of short range echoes are completely explained.

[2] A. H. Taylor and L. C. Young, Proc. I. R. E., **16**, 561; May, 1928 J. B. Hoag and V. J. Andrew, Proc. I. R. E., **16**, 1368; October, 1928.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

DETECTION CHARACTERISTICS OF SCREEN-GRID AND SPACE-CHARGE-GRID TUBES*

By
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and
Birney Dysart
(Stanford University, California)

Summary—*This paper reports the results of measurements made to determine the grid-leak grid-condenser detection properties of screen-grid and space-charge-grid tubes. It is found that the rectifying properties of the grid circuit for the four-element tube used in these ways is about the same in character and magnitude as in three-element tubes with the same type of filament. The rectifying action in the grid circuit is largely independent of the voltages of the filament, plate, and second grid when compared at the same grid resistance. The space-charge-grid leak-condenser detector is superior to the screen-grid and most three-element grid leak-condenser detectors in that the space-charge-grid tube retains its full rectifying powers at adjustments which give full reproduction of the high notes. Grid detection with four-element tubes is superior or inferior to grid detection with three-element tubes primarily to the extent that audio-frequency amplification is more or less satisfactory in the two cases.*

Introduction

The purpose of this paper is to give data on the grid-leak detection characteristics of four-element tubes, a continuation of a similar study made of three-element tubes[1]. The present results are expressed in terms of the detector voltage constant v as a function of grid resistance, for reasons that are explained in the original study,[1] and are obtained from bridge measurements[2] of dynamic grid resistance at grid voltages that differ slightly.

Performance

Screen-Grid Tube.— Fig. 1 (a) shows the connection for grid-leak grid-condenser detection with a screen-grid tube. The inner grid (G in the figure) is used here as the control grid. The rectification resulting from this connection is represented by Fig. 2, which shows the general type of v-R_g curve obtained from the screen-grid tube, while holding E_{sg} and E_p constant. The value of v rises rapidly for values of grid resistance less than 200,000 ohms. So long as the operating point is such that R_g is greater than 200,000 ohms, v has a value which may be considered to be fixed at -0.4.

* Dewey decimal classification: R 134. Original manuscript received by the Institute, December 3, 1928.

[1] F. E. Terman and Thomas M. Googin, "Detection Characteristics of Three Element Vacuum Tubes," Proc. I.R.E., 17, 149; January, 1929.

[2] F. E. Terman, "Some Principles of Grid-Leak Grid-Condenser Detection," Proc. I.R.E., 16, 1384; October, 1928.

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Variation of E_p, E_{sg}, and E_f has only a small effect on the value of v even when on negative resistance parts of the characteristic. Thus the minor influence which the screen-grid potential has in de-

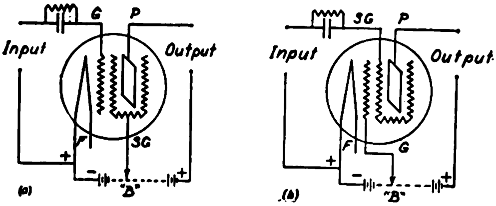

Fig. 1—Connection for Grid-Leak Grid-Condenser Detection with (a) screen-grid tube, and (b) space-charge-grid tube.

termining the detection characteristics is shown by Fig. 2. Variations in v resulting from a change in E_p are small and lie within the allowance for possible experimental error. Besides changing slightly with changes

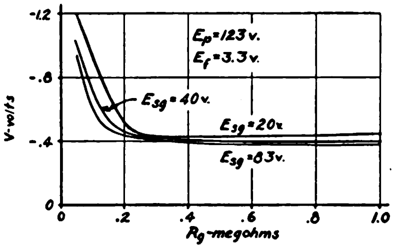

Fig. 2—Typical $v-R_g$ Characteristics of Screen-Grid Tube.

in E_p, E_{sg}, and E_f, it has been found that v may differ a little for individual tubes as is witnessed by Fig. 3.

Space-Charge-Grid Tube.— Fig. 1(b) shows the connection for grid-leak grid-condenser detection with a space-charge-grid tube.

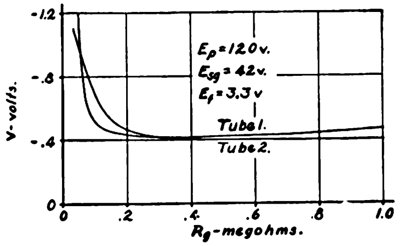

Fig. 3—Comparison of Detection Charactistics of Two Screen-Grid Tubes.

Here the outer grid (*SG* in the figure) is used as the control grid. The rectification resulting from this connection is represented by Fig. 4, which shows the general type of v-R_g curve obtained from the space-

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charge-grid tube, when E_g and E_p are held constant. For the whole range of R_g that was explored (50,000 to 2,000,000 ohms), v was practically constant at -0.4.

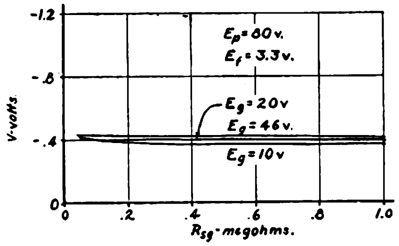

Fig. 4—Typical $v-R_g$ Characteristics of Space-Charge-Grid Tube.

The variations of v resulting from changes in E_p, E_g, and E_f are again of the same order of magnitude as for the screen-grid tube. Thus a change in E_p has no apparent effect on v, while the space-charge-grid potential is only of secondary importance as Fig. 4 shows. The value of v also may differ a little for individual tubes, as is shown by Fig. 5.

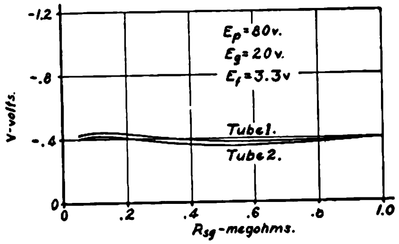

Fig. 5—Comparison of Detection Characteristics of Two Space-Charge-Grid Tubes.

Discussion

The results of the investigation of the grid-leak grid-condenser detection characteristics of the screen-grid and space-charge-grid tubes are very similar in character and magnitude to those results obtained for three-element tubes with the same type of filaments. The 201A type of tube with its thoriated tungsten filament is an example of such a three-element tube and in fact the 201A tube has a voltage constant slightly inferior to the voltage constant of the four-element tube now on the market.

In the case of screen-grid detection, the v-R_g characteristic curve rises rapidly above -0.4 for values of R_g less than 200,000 to 300,000 ohms; while the v-R_g characteristic curve for space-charge-grid detection remains constant at -0.4 to values of R_g less than 50,000 ohms.

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(See Figs. 2 and 4.) This means that the space-charge-grid tube is superior to the screen-grid tube as a detector because it enables the high notes to be reproduced with much better quality without sacrificing sensitivity.

AVERAGE DETECTION CHARACTERISTICS OF A FEW TUBE TYPES

Type	v (Flat part)	R_g at start of flat part (approximate)	μ	Highest undistorted frequency at full sensitivity*
201A	−0.47	150,000	8	3,500
227	−0.23	50,000	8	11,500
322(SG)	−0.40	250,000	300	2,200
322(SCG)	−0.40	Less than 50,000	60	Over 11,500

* For an effective grid condenser capacity of 0.0003μf.

The above table shows a comparison of the detection characteristics of several common tube types, including the screen-grid and space-charge-grid tubes, and is to be compared with Table I of the previous paper.[1]

USES OF FOUR-ELEMENT DETECTORS

Space-Charge-Grid Detector.— The space-charge-grid type of detector has the average grid rectifying action of the usual three-element tubes plus the advantage of space-charge-grid audio-frequency amplification, which is unexcelled for resistance or impedance coupling. In addition excellent quality may be secured at the highest audible frequencies without a sacrifice of rectifying properties.

Screen-Grid Detector.— The screen-grid type of detector also has about the same rectifying action on the grid side as the average tube. When systems are devised for making use of the tremendous amplification of the screen-grid tube at audio-frequencies this tube will be an excellent detector, although having the disadvantage of requiring a very small grid condenser if the high audio notes are to be reproduced with good quality and maximum sensitivity.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

A DIRECT READING FREQUENCY BRIDGE FOR THE AUDIO RANGE, BASED ON HAY'S BRIDGE CIRCUIT*

By
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Summary—*This paper describes the construction and operation of a direct reading bridge for measuring audio frequencies in the range from 50 to 5000 cycles per second and upwards. The circuit used is a modification of Hay's bridge for the measurement of inductance, and is obtained from it by the insertion of a variable resistance in the inductive arm. This resistance and another in the condenser arm are varied for balance. For the design as given, the frequency in cycles per second is determined by the number of ohms in the variable resistance plus that of the internal resistance of the coil in series with it.*

The bridge is portable, simple to construct, and makes use of external connection for the variable resistances to two standard decade non-inductive resistance boxes. Its operation is convenient and permits of a precision of balance of 0.1 per cent with a probable error of less than about 0.25 per cent.

THERE are several bridge circuits having reactive arms in which the balance is dependent upon frequency; but not all of these are suitable for a direct reading frequency bridge having a long range, absorbing negligible power, and offering facility of operation and construction.

In Hay's bridge[1] for measuring inductance, the equation for the inductance is nearly independent of frequency, but the equation for the resistance of the inductor is dependent upon frequency and a variable resistance.

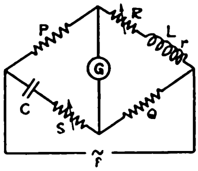

Fig. 1—P, Q—fixed non-inductive resistances; S—variable non-in-ductive resistances; R—variable non-inductive resistances (includes resistance r of coil); C—mica condenser.

This suggests a modification of the circuit by introducing a variable resistance in series with the inductance, as shown in Fig. 1.

For null balance in the detector:

$$PQ = (R + L\omega j)\left(S + \frac{1}{C\omega j}\right). \qquad (1)$$

* Dewey decimal classification: R 201.6. Original manuscript received by the Institute, January 17, 1929. Presented before Toronto section, October 10, 1928.

[1] C. E. Hay, Inst. P.O. Engrs., November, 1912.

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By equating real and unreal parts:

$$PQ=\frac{L}{C}+RS \tag{2}$$

and

$$\frac{R}{C\omega j}+LS\omega j=0. \tag{3}$$

The two conditions given by (2) and (3) must be satisfied for balance of the bridge. Equation (2) signifies that, since P, Q, L, and C are constant in value, the product of R and S for any balance must be equal to a certain constant.

It may be shown mathematically and is proved experimentally that the two balances by adjustment of R and S are rapidly convergent and nearly independent of one another.

From (3).

$$\omega^2=\frac{R}{CLS}\cdot \tag{4}$$

Substituting from (2) to eliminate S:

$$\omega^2=\frac{R^2}{CL\left(PQ-\frac{L}{C}\right)} \tag{5}$$

whence

$$=\frac{R}{2\pi\sqrt{L(PQC-L)}}=R\times\text{constant}. \tag{6}$$

Design

Equation (6) shows that the bridge may be so designed that the frequency is determined directly in terms of a single variable resistance, or by a convenient multiple of the number of ohms thereof..

For the use required of this bridge it was decided to make it direct reading, and to have maximum sensitivity at 1000 cycles per second. This requires that:

$$2\pi\sqrt{L(PQC-L)}=1 \tag{7}$$

and that $L=1/2\pi$ henry, and $C=1/2\pi$ microfarad.

This choice of inductance and capacitance will make the phase angle of the reactive arms 45 degrees at 1000 cycles pers second. This is the condition for which a small change in either resistance or reactance will produce maximum effect upon the bridge balance.

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To satisfy (7) the product of P and Q must be 2×10^6, and so that the impedance of all four arms shall be equal at 1000 cycles per second, both P and Q are chosen equal to $\sqrt{2}\times10^3$ ohms.

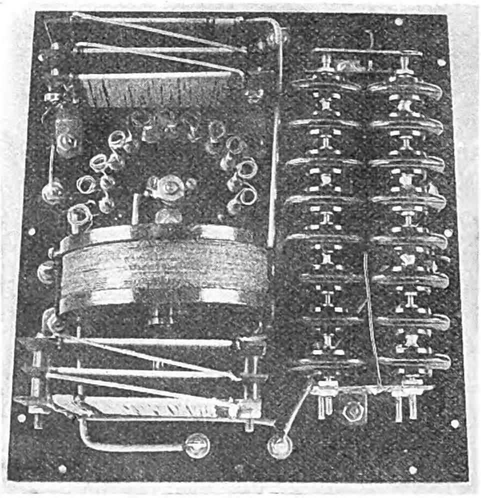

Fig. 2

Construction

Fig. 2 shows the construction employed, and Fig. 3 shows the instrument (with cover removed) connected to two external non-

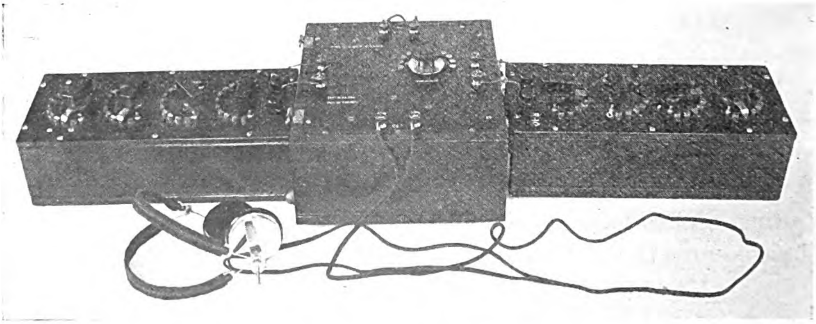

Fig. 3

inductive decade resistance boxes. The range of the instrument is from 50 cycles per second upwards, 5000 cycles per second, for practical reasons, being about the upper limit for convenient aural balance using telephones.

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Below 1000 cycles per second it is convenient to use for the variable resistance R, a four-dial decade resistance box calibrated in hundreds, tens, units, and tenths, and for the variable resistance S a four-dial resistance box calibrated in thousands, hundreds, tens, and units. Below 100 cycles per second a fixed 10,000-ohm resistance can be connected in series with the S resistance box. For frequencies above 1000 cycles per second it is convenient to interchange the two resistance boxes.

Both the P and Q resistances are wound of No. 39 (B. & S.) S. C. manganin wire on three bakelite cards apiece, each 4 in.×2 1/2 in.× 1/8 in. The winding is made in Ayrton-Perry fashion so as to be non-inductive.

The inductance coil L is wound of No. 26 (B. & S.) enamelled wire with 1465 turns. It is wound on a spool 2 1/8 in. in diameter and 3/4 in. wide, and the outside diameter is 3 5/8 in. Glassine paper 0.001 in. thick is used between layers. The resistance of this coil at 20 deg. C. measures approximately 45 ohms.

Since the resistance r of the inductance coil must be included in the resistance R, a bifilar series resistance of manganin wire is added to make r the convenient value of 50 ohms. This value is added to the resistance in the external resistance box at balance and the total number of ohms gives the frequency in cycles per second. Part of this additional resistance is connected between the contacts of a rotary switch to compensate in steps of 1 deg. C. for variations of the coil resistance with temperature between 15 deg. C. and 25 deg. C. This extra adjustment is a refinement which is necessary only for precise measurements and particularly for frequencies below 1000 cycles per second.

If the bridge were constructed complete with self contained variable resistances, the necessity of adding 50 ohms to the dial reading could be avoided by making the zero of the dials (of R) begin with 50 ohms, i.e., 50 cycles.

The distributed capacitance of this coil measures approximately $40\mu\mu f$. Since the variable part of the resistance R is increased with frequency, the slight increase in the coil resistance r due to capacity and eddy-current effects will be negligible.

A single high grade mica condenser of the value required could not be obtained readily, so sixteen $0.01\mu f$ units of a well known moulded type, matched to give the desired capacitance, were connected in parallel. As shown by Curtis[2] and others,[3] the variation in

[2] Curtis, Bulletin Bureau of Standards S-137.
[3] Dictionary of Applied Physics, Glazebrook, Vol. II, p. 139.

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capacitance of a high grade mica condenser with temperature and frequency is small, i.e., of the order of 0.1 per cent over the audio range.

Small differences in the values of *L* and *C* from the theoretical values previously given are readily corrected for by altering either the *P* or *Q* resistance.[4]

Operation

It is undesirable that any ground should be applied to the bridge. If the a.c. source be grounded the insertion of a suitable transformer between it and the bridge is recommended. However, if the ground cannot be avoided least error will occur for the connection of the grounded line to the *L* and *Q* junction. If an amplifier be used to precede the detector a grounded filament connection should not be applied directly to the bridge.

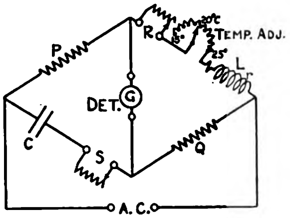

Fig. 4—Circuit of the Actual Instrument.

For very precise measurements a Wagner Earth with arms similar to the *C S* and *Q* arms and grounded at their junction may be connected across the source. This is adjusted in the usual manner after a preliminary balance of the bridge.

With careful balancing of the *R* and *S* adjustments a precision of adjustment in the *R* balance of 0.1 per cent can be obtained over a range of 50 to 5000 cycles per second. Due to possible variations in the capacitance of the condenser *C* and the limit of adjustment accuracy in the resistance boxes, and allowing for the balancing error, the probable error will be within 25 per cent. The precision and the accuracy of the balance will, of course, depend upon the detector used.

Between 200 and 2000 or 3000 cycles per second, telephones are satisfactory as a detector, and for ordinary work may be used directly with only a few volts applied to the bridge. For very accurate work or where the voltage of the unknown emf is low, or on the extreme ends

[4] After the preparation of this paper and the experimental development were complete, it was discovered that a bridge operating on the same principle but lacking some of the features of the authors' development and using different circuit constants had been described previously in a Japanese paper: K. Kurokawa and T. Hoashi: *Jnl.* I.E.E., Japan, No. 437, pp. 1132–1138.

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of the frequency range, a two-stage amplifier of good quality preceding the detector is useful.

Harmonics in the waveform being investigated are strongly in evidence when the bridge is balanced. At the very low frequencies, especially where the harmonics are much more audible than the fundamental they make it difficult to obtain a precise balance. The use of a good low-pass filter to eliminate harmonics is recommended.

To obtain the precision claimed it is necessary, below about 200 cycles per second, to make use of a vibration galvanometer as the detector, due to the insensitiveness of the ear and the telephones. Since this is a resonant device little trouble from harmonics is experienced. Given a pure waveform an amplifier and rectifying vacuum tube with microammeter, as recommended by Turner,[5] can be used to advantage as a visual form of detector.

Effect of Variation of L and C

Since the equation of the bridge used is $f=R$, it is desirable to know how R will vary with C, L, P, or Q, as errors in the original adjustment, or subsequent variation of these quantities, may affect the accuracy of balance.

From (2) and (4):

$$R=\omega\sqrt{L(PQC-L)} \tag{8}$$

$$S=\frac{1}{\omega C}\sqrt{\frac{PQC-L}{L}}\,. \tag{9}$$

Differentiation of (8) with respect to the several variables shows that a small change in P, Q, or C causes the same percentage change, in the same sense, in the value of R. However, for the particular choice of constants employed in this design which makes $(2L\text{-}PQC)=0$, this process shows that small changes in the value of L do not affect the R balance. This result is advantageous since errors in the measurement of the inductance L or variations in it are compensated for.

Differentiation of (9) with respect to L discloses the fact that a small change in the value of L affects the bridge by requiring the same percentage change, in the opposite sense, in the value of S.

To verify the last two statements experimentally, increments in inductance up to 3 per cent were added in series with the coil L, and to maintain balance S had to be decreased in the proportion predicted, while there occurred negligible change in the R adjustment.

[5] H. M. Turner, Bulletin of A. I. E. E. Symposium on High Frequency Measurements, May, 1927.

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Differentiation of (9) with respect to C shows that the S adjustment is independent of small changes in the value of C. Since the capacitance C is likely to vary more in value with changes in temperature and frequency than the inductance, it might appear that the frequency should be determined preferably in terms of the resistance S. However, this arrangement would not be direct reading, and S must include the unknown resistance of the condenser. Fortunately the resistance of such a condenser varies approximately inversely as the frequency and likewise for the value of the S adjustment. Thus the resistance of a good mica condenser of the value employed is negligible at any frequency compared to the value of the resistance S required for balance.

In another report[6] further details of the operation of the bridge, especially for the less audible frequencies, are given. Also, its conversion by means of a double-pole double-throw switch for use as an ordinary Hay's bridge for measurement of large inductances is described.

[6] Bulletin School of Engineering Research, University of Toronto, 1929.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

VOLTAGE SURGES IN AUDIO-FREQUENCY APPARATUS*

By
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Summary—*Transient voltages of over 2000 are shown to occur across the secondary when the normal plate current of a high inductance audio transformer is opened. The oscillograph and inverted vacuum tube are used to bring out further that these transients are oscillations of definite frequency and magnitude, depending on the primary current, inductance, and secondary distributed capacity. The manner of breaking the circuit has a minor influence, for in most cases no appreciable discharge takes place at the contacts.*

The effect of amplifier circuits on the transients is considered, and suggestions are made as to possible means of protection.

Introduction

It is common knowledge that if the current through an inductance is broken, the collapse of the magnetic field may result in a tremendously high voltage in the windings. Audio-frequency transformers show no exception, and the magnitude and characteristic nature of their transients, as well as the effect of associated apparatus, make a subject well worthy of special study.

Consider a typical amplifier circuit such as Fig. 1. If the plate circuit of the first tube is opened at X there will be a sudden rise of

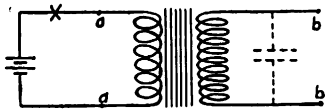

Fig. 1—Amplifier Circuit on Which Transients Were Measured.

voltage across the primary. A similar rise, greater by the turn ratio, will take place in the secondary.

If this secondary voltage is of such polarity as to make the grid of the following tube positive, a current will flow through the circuit and kill the surge. Should the grid be made negative, however, no current could flow from the transformer terminals and the voltage would be free to rise.

Under this latter condition the true nature of the transient is shown, and it is found that the distributed capacity is a determining factor. The voltage will increase to a value such that the magnetic energy formerly stored in the iron core is transformed into electro-

* Dewey decimal classification: R 390. Original manuscript received by the Institute, November 20, 1928.

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static energy in the distributed capacity. This means that we have the ordinary parallel resonant circuit, as indicated in Fig. 1, with a perfectly definite maximum voltage. The surge is therefore the start of a highly damped oscillation in the secondary, but as soon as the voltage drops to zero and starts to rise in the opposite direction, the grid goes positive and kills the remainder of the oscillation.

In the second tube, when a negative surge reaches the grid, the plate current is stopped as effectively as if the circuit had been snapped open. There is no possibility of a loss of energy at the contacts, and the transient voltage is limited only by the distributed capacity. If this second transformer is of the modern, high inductance type, a transient voltage of over 2000 across the secondary is not uncommon.

METHOD OF MEASURING TRANSIENTS

The measurement of these transients calls for a means which is rapid and at the same time uses practically no energy. These requirements are adequately met by the oscillograph and inverted vacuum tube[1], arranged as in Fig. 2.

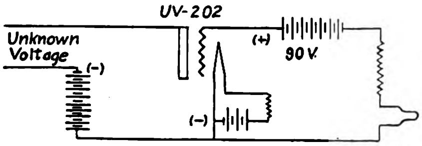

Fig. 2—Inverted Tube and Oscillograph Circuit.

In this the functions of the grid and plate are reversed and the grid is made positive, thus drawing a current sufficient to operate an oscillograph element. Any negative voltage applied to the plate then has an effect on the grid current equivalent to reducing the grid voltage by $1/\mu$ times as much as that of the plate.

A five-watt tube was used, with a grid battery of 90 volts, and sufficient resistance in the circuit to limit the grid current to 30 ma at zero plate voltage. These values gave a plate voltage-grid current curve that was quite straight over the greater part of the range. See Fig. 3.

Since the transients were oscillatory, it was necessary to provide a negative plate bias as shown in Fig. 2, to keep the plate from going positive when the oscillation reversed.

In all oscillograms the circuit was broken with an ordinary telegraph relay. When the relay was slowed down so as to barely open, any

[1] Terman, "The Inverted Vacuum Tube", PROC. I.R.E., 16, 447; April, 1928.

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transient that would have gone above 200 volts or so, in the first transformer primary, was limited to that value by a discharge at the contacts. However, with normal operation of the relay, or upon the fairly rapid opening of a knife switch, the gap was apparently open before the voltage had time to build up to any appreciable amount.

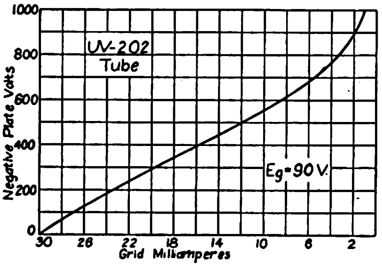

Fig. 3—Calibration Curve of the Measuring Apparatus.

Characteristics of the Transients

The nature of the transients is best shown by considering the transformers apart from the amplifier circuit. Oscillogram No. 1 (Fig.4) is typical of the results. The line just below the 60-cycle timing wave

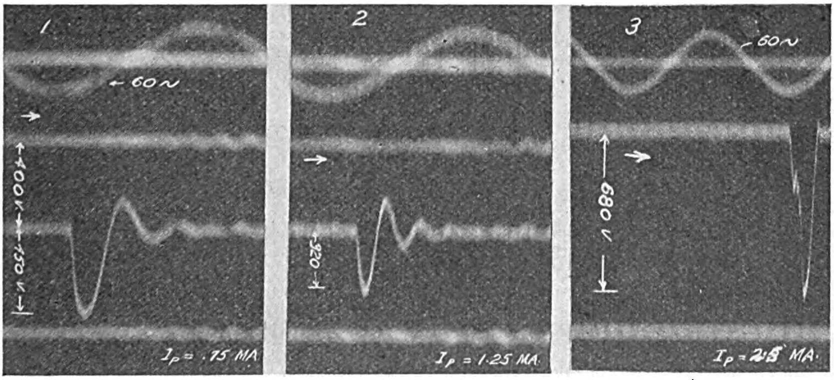

Fig. 4—Transients from a Transformer Not Connected in the Amplifier Circuit.

shows a grid current of 30 ma, i.e., zero plate voltage. In the middle line, in which the transient occurs, the grid current is reduced by a 400-volt plate bias. The bottom line shows zero grid current.

For this transient a current of 0.75 ma, as produced by one dry cell, was broken in the primary of a 3.5:1 transformer. The trans-

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former had a primary inductance of 106 henries and was manufactured by the Ferranti Co. The secondary voltage (across *b,b*, Fig. 5) reached a value of 450 as shown. With a 6:1 transformer of 13 henries primary inductance, manufactured by the General Radio Co., the breaking of 1.25 ma gave a transient of 320 volts on the secondary (oscillogram No. 2).

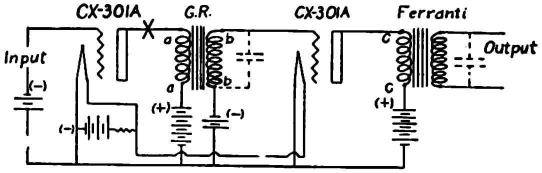

Fig. 5—Single Transformer as Used for Oscillograms 1, 2, and 3.

When the current in the primary of the Ferranti transformer was increased to the not uncommon value of 2.5 ma and the circuit opened, the surge could be measured only in the primary, and even for this the plate bias had to be removed. The voltage reached 680 (No. 3), which meant over 2000 in the secondary. The oscillation was killed as soon as the plate voltage of the inverted tube dropped to zero and started to go positive.

As already pointed out, the oscillation is apparently due to resonance between secondary inductance and distributed capacity. If this is so, it should be possible to determine the transformer constants and calculate the transient voltage.

The original inductive energy is readily obtainable from the relation

$$W = \tfrac{1}{2} L_p I_p^2. \tag{1}$$

At the peak of the voltage wave this inductive energy has been transformed into electrostatic energy so that

$$W = \tfrac{1}{2} C_s E_s^2. \tag{2}$$

Combining the two, and solving for secondary voltage

$$E_s = \sqrt{\frac{L_p}{C_s}}\, I_p. \tag{3}$$

The secondary distributed capacity may be determined from the frequency of oscillation on the oscillograms, according to the relation

$$C_s = \frac{1}{(2\pi f)^2 L_s} \cdot \tag{4}$$

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Losses are not considered, as the accuracy of the results does not warrant it. The voltages thus calculated, while of the same order of magnitude, show a variation of as much as 35 per cent on either side of the measured value. A more accurate determination of the distributed capacity would cut down the error somewhat, but it must be observed that the secondary will show a varying effective capacity, depending on the leads and tube connected to it. The above agreement is therefore sufficiently close for present purposes.

The results of tests on widely different types of apparatus are given in Table I, and show the general magnitude of transients to be expected.

TABLE I

OBSERVED AND COMPUTED VALUES OF TRANSIENTS CAUSED BY OPENING THE PRIMARY CIRCUIT

I_p ma	Circuit Tested	Inductance of Circuit henries	Freq. of Surge ~/sec.	Sec. Dist. Cap. (Calc.) $\mu\mu f$	Crest Volts (Calc.)	Crest Volts (Obs.)	Crest V. / $I_p\sqrt{L}$
0.50	Sec. Ferranti 3.5:1	1410	..	..	..	312	16.5
0.75		1300	..	..	..	450	16.7
1.00		1220	..	..	..	630	18.0
1.25		1160	300	213	550	850	20.0
0.50	Prim. Ferranti 3.5:1	115	..	..	..	115	21
1.00		100	..	..	..	210	21
1.25		95	..	..	..	365	30
1.75		84	..	..	..	530	33
2.50		76	..	..	..	640	29
1.25	Sec. G.R.6:1	470	500	216	215	320	11.8
2.50		465	..	..	..	650	12.0
3.0	Sec. R.C.A. 9:1	118	2000	54	350	265	8.1
6.8		118	..	..	..	625	8.5
2.5	Phones W.E. 2200 ohm	4.5	2000	1480	100	75	14

For purposes of comparison, all voltages are divided by I_p and $\sqrt{L}$, and in that form a fair regularity is shown. The values of distributed capacity are about as could be expected, considering the concentration and physical size of the windings. It is also interesting to note that the transient voltage per ma is greater for the higher values of current; a point for which no explanation is at present apparent.

The Ferranti transformer had a small bypass condenser built in across the primary, but the fundamental wave of the transient showed no change when this was disconnected. However, oscillograms Nos. 3, 9, and 10 show small, high-frequency oscillations superimposed on the main wave which may be due to resonance between primary distributed capacity and inductance. These should be affected by the additional primary capacity, but were too irregular for a careful study.

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Transients in Amplifier Circuits

The effect of the amplifier circuit on these transients is of considerable importance. Oscillogram No. 4 was taken across the points *a,a* of Fig. 1, and is the same as No. 2, except that the primary current (1.25 ma) was broken through the plate circuit and the secondary was connected so as to make the grid of the following tube negative. The additional tube capacity across the secondary makes the seecondary voltage about 10 per cent less than in No. 2, showing 290 volts across the grid-filament of the second tube. The frequency of oscillation is correspondingly a little lower, and the latter part is suppressed when the grid of the amplifier tube goes positive.

The effect on the following primary (points *c,c*, Fig. 1) is shown in oscillogram No. 5. The highly negative grid stopped the plate current as rapidly and effectively as if the circuit had been physically opened. The transient is of the same magnitude, per ma of current broken, as No. 3. With higher currents, the negative grid is even more effective for there is no possibility of a discharge at the contacts. The voltage of 530, multiplied by the turn ratio, represents the amount that is applied to the grid of the following tube.

Reversing the secondary of the first transformer kills the transient caused by opening the circuit. However, the closing then causes a disturbance in the secondary as shown in No. 6, which results in No. 7 in the following primary.

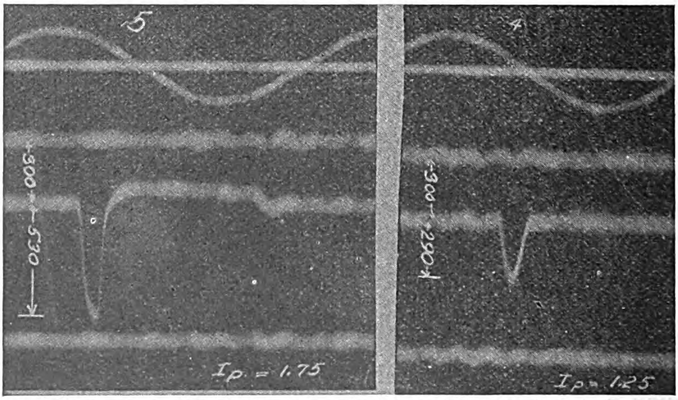

Fig. 6—Transients from Opening the Plate Circuit of an Amplifier.

These illustrations were taken with separate plate batteries in order to isolate the different effects. With a single battery, and the transformers connected according to the manufacturers' terminal markings, the transient No. 8 occurred across the primary of the second

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transformer, upon opening the common battery lead. It is a combination of several transients, and shows the same maximum value, per ma of current broken, as No. 3, taken across the same primary alone. When an attempt was made to reverse the connections on the secondaries, oscillation took place.

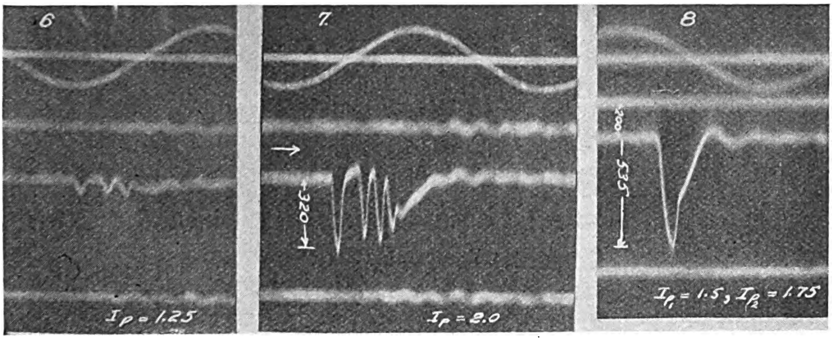

Fig. 7—Surges Caused by Closing Plate Circuit (Nos. 6 and 7) and by Opening a Common Plate Battery Lead (No. 8).

The current for tests on the amplifier circuit was obtained through a 201A-type tube without departing from recommended values of grid and plate voltage, and the secondary surges approached 2500 volts. This voltage would jump a gap of 3/16 in. The use of larger tubes, now quite common, makes possible even higher transients in coupling and output transformers.

One may say that the ordinary radio set does not have the plate circuit opened, except perhaps in testing. However, a small surge as

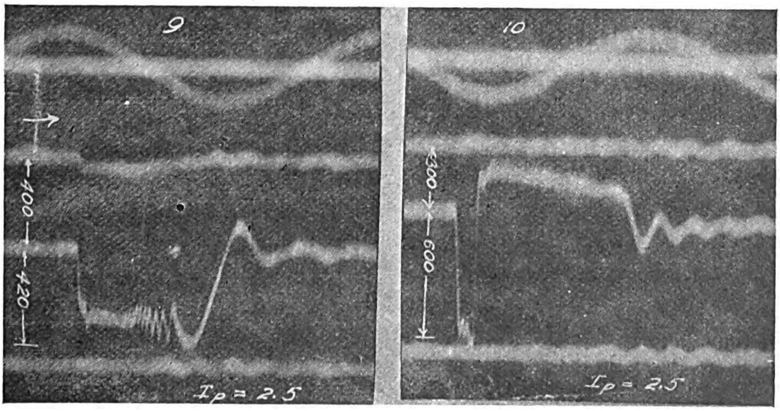

Fig. 8—Secondary Discharges across a Spark Gap.

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the "B" eliminator voltage built up might be equivalent to opening the circuit in the last stage. Pulling a tube out of the socket with the filament on, or plugging speakers in, might easily cause transients of the magnitudes observed.

The transformers used for the oscillograms were well built and stood the stress, but one of a different make was found which apparently flashed across internally, thus giving an indication of why transformers sometimes "go noisy."

Protection Against Transients

A method of protecting transformers which appears both simple and effective would be to incorporate a small safety gap across the secondary. Oscillogram No. 9 shows a secondary transient that would normally have gone over 2000 volts, limited to 420 by a gap of the "horn" type, made of No. 18 wires and separated by a thickness of tissue paper. In this test the primary was connected direct to a 4 1/2-volt battery. In No. 10 the same current was broken with the gap slightly wider and the transformer connected back in the second stage of the amplifier circuit of Fig. 1.

As to the type of gap, a pair of points would have the least capacitance, but their appreciably slower action makes them undesirable. If the capacitance could be kept low enough, a small pair of plates separated by a perforated mica strip would be most effective.

Conclusion

The tests of this paper show that the transients in audio-frequency apparatus may go to seriously high values. Also they are of definite value and dependent on determinable factors. After further investigation, one should be able to take a few relatively simple measurements and predict quite accurately just what maximum voltage could be expected.

This gives a direct indication of insulation requirements. It might be further added that when the time comes that the radio industry shall be ready to adopt a set of standardization rules on the subject, there will be a definite foundation on which to work.

The writer wishes to express his sincere thanks to Dr. Frederick E. Terman, of the Stanford University faculty, for his valuable guidance and suggestions during the course of this investigation.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

COLD CATHODE RECTIFICATION*

By
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Summary—*Asymmetric conductivity in a gas between two electrodes can be accomplished without the use of a hot cathode, if the relative areas of the electrodes are widely different. This paper presents the results of an investigation of the discharge phenomena and asymmetric conductivity in a cold cathode rectifier tube containing two anodes and one cathode, when utilizing: (1) one anode and one cathode; (2) two anodes and one cathode; and (3) two anodes and one cathode and a low-pass filter circuit. The electrodes are situated in a gaseous atmosphere of helium and neon at 20 mm Hg. pressure.*

A theoretical explanation of the asymmetric conductivity is considered; (1) for the initial state of asymmetry; and (2) for the limiting state of asymmetry, the former referring to the phenomena of ionization by collision and the latter to the theory of normal cathode current densities.

Typical static characteristic curves are shown, from which can be anticipated the asymmetry shown by the oscillographic records.

Theoretical expressions are given for degree of asymmetry in half- and full-wave rectification, and data collected with reference to the performance of the rectifiers show that the theoretical values are closely approached by this cold cathode type of tube.

Introduction

THE subject of asymmetric conductivity in gases between physically dissimilar electrodes goes back many years. The bibliography at the end of this paper gives specific references to some of the early literature in this field. Of the more recent investigations, the work of Kneser,[1] Jolley,[2] Graetz,[3] and Günther-Schulze[4] is of particular interest. These investigators describe rectification devices which utilize one anode and one cathode, thus giving one-way asymmetry. Charlton[5] has secured design patents covering cold cathode devices possessing two anodes and one cathode, which thus give two-way asymmetry. The patents describe electric discharge devices of small and large current capacity, the capacity being determined by the surfaces available for heat radiation. One form of tube developed by Charlton is capable of handling effective potentials as

* Dewey decimal classification: R 149. Original manuscript received by the Institute, January 28, 1929.

[1] *Annalen der Physik*, **72**, 519–24; 1923.
[2] "Alternating Current Rectification," 2nd Ed., p. 319; 1927.
[3] "Die Elektrizität," pp. 511–13, 1928.
[4] "Elektrische Gleichrichter und Ventile," 1st Ed., pp. 32, 130, 1924.
[5] British Patent Specifications 237, 235 and 237, 236, July 17 1924.

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high as 2000 volts and currents of the order of several hundred milliamperes, with quite complete rectification.

This paper is the result of a more complete study of the influence, of variation in load, upon the voltage and current waves of a cold cathode rectifier tube with particular reference to the reverse current.

Theory

Cold cathode devices for rectification are essentially identical in that they embody a cathode, of relatively greater surface area than the anode, situated in a suitable gas of the proper density to produce the desired conductivity by ionization.

In considering the initial state of asymmetry i.e., when the first rectifying discharge occurs, we may recall a theory expressed by Townsend[6] in explaining the asymmetry he obtained in various gases, at low pressures, between two electrodes, one in the form of a tube and the other a rod coincident with the axis of the tube. Townsend attributed the unidirectional conductivity he obtained to the ionization

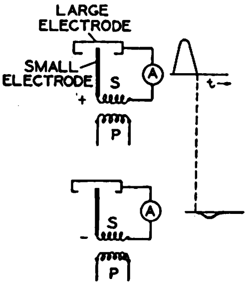

Fig. 1—Current Cycle in One-way Rectification.

by collision arising in the more intense electrostatic field near the rod or smaller electrode. This explanation, however, seems to be inadequate to account for the experimental fact to be discussed later, namely, that with a given pair of dissimilar electrodes, the initial direction of the rectifying discharge is not always dependent upon the polarity of the electrodes even though the ultimate direction is fixed. The mechanism of the initial state is not very well understood at the present time and whatever it may be, any theoretical explanation of it will have to take account not only of the nature of the gas used but also of the nature of the material[7] of the electrodes, a point not considered in Townsend's original theory. The next part of the theoretical discussion concerned with the limiting state agrees quite well with

[6] Townsend, Ionization of Gases by Collision, pp. 11–15.
[7] Holst and Osterhuis, *Phil. Mag.*, **46**, 1117; 1923.

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experimental fact, but it is restricted to currents that exceed the normal cathode current density, whereas a theory of the preceding state will have to explain the situation for currents less than or at most equal to the normal value.

In considering now the limiting state for unidirectional conductivity, we have recourse to the theory of normal cathode current densities and fall of potential. In regarding the operation of the rectifier having one large electrode and one small electrode, let us take the case of platinum electrodes in helium at 20 mm Hg. pressure.

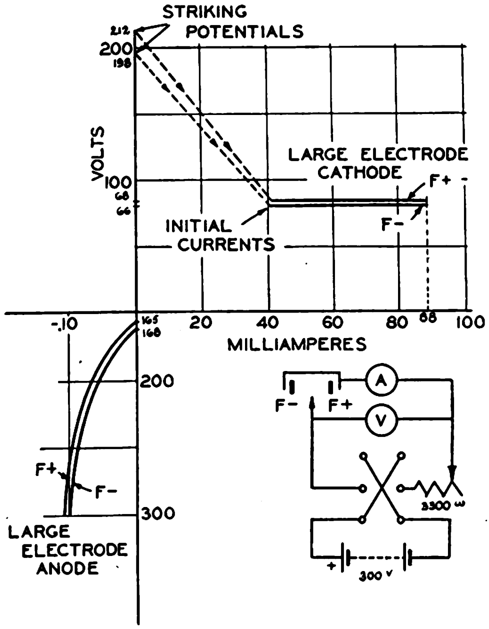

Fig. 2—D.C. Characteristic Curves.

The normal current density for a platinum cathode in helium[8] has been observed up to 4 mm Hg. pressure, and with the normal current density assumed approximately proportional to the square of the pressure, a current of 4.0 milliamperes per square centimeter is available at 20 mm Hg. pressure. For different cathode materials the normal current density is of the same order of magnitude as it is for platinum.

Let us now consider that alternation of the cycle during which the large electrode is cathode. During this interval, for increasing current values, the current per square centimeter of the cathode and

[8] Bär, *Handbuch der Physik* (Springer), Vol. 14, p. 211.

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the cathode fall of potential remain practically constant, until the current becomes so large that the entire cathode surface is covered with the cathode glow. When this stage is reached the normal current density increases but is accompanied by a very great rise in the potential. With a normal current density of 4.0 milliamperes, a cathode having an area of 22.0 square centimeters will allow a current of 88.0 milliamperes to flow, with normal fall of potential. It is during this alternation in the cycle, i.e., when the large electrode is functioning as cathode, that the greatest current flow occurs and makes the large electrode positive with reference to the external circuit.

When the small electrode, with an area of 7.0 square millimeters, becomes cathode, during the next alternation in the cycle, the normal current stage ceases when 0.28 milliampere flows. A great increase of potential is required to force currents greater than this through the

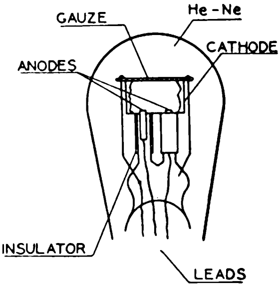

Fig. 3—Construction of Tube.

tube. In other words, the tube acquires a high resistance so that the maximum voltage applied to it causes a much smaller current to flow than when the large electrode is functioning as cathode. In the ideal case there is a complete suppression of the negative current, as we shall see later. This cycle of events is depicted in Fig. 1.

The static characteristic curves given in Fig. 2 for one tube enable us to predict this difference in current flow, depending upon whether the small or large electrode is cathode. Further interpretation of these curves will be given in connection with one way rectification.

In two-way asymmetry, using two small electrodes, the discharge is maintained alternately between the large electrode and either small electrode. This results in full-wave rectification.

Apparatus

A tube constructed as shown in Fig. 3 was used for the experiments. The electrodes are made of nickel and the tube is filled with a gaseous

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mixture of helium and neon[9] at 20 mm Hg. pressure. In general, for rectification purposes, the pressure of the gas in the tube may vary between 5–80 mm Hg. depending upon the characteristics of the gas which is employed, the operating potential, the current carrying capacity, and other tube conditions. Charlton[10] found that for helium a pressure[11] of 19 mm Hg. gave a minimum voltage drop between the electrodes he used. By introducing into the tube an alkali metal such

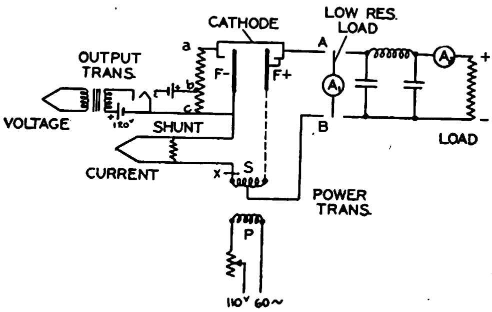

Fig. 4—Connection of Apparatus.

as caesium the voltage drop can be lowered still further. In the construction of the tube, the two small electrodes are placed sufficiently far apart or else placed on either side of the cathode which may be in the form of a disk. This prevents a discharge from taking place between the two small electrodes which would eventually change to an arc discharge.

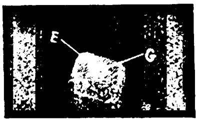

Fig. 5—Initial Glow Discharge. E is glow between $F-$ electrode and cathode, G is glow through gauze.

The connection of the apparatus is given in Fig. 4, the tube being joined to a transformer, the voltage of which is varied over the range desired.

The voltage and current across the terminals of the tube were recorded photographically by means of a Duddell oscillograph. The

[9] Neon has a low cathode fall, hence a small starting potential.
[10] Loc. cit.
[11] What influence the pressure has upon the voltage and current curves has yet to be determined. This investigation was carried out at a constant pressure of 20 mm Hg.

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voltage variations were taken from a part bc of a high resistance ac connected in parallel with two electrodes, i.e., a large electrode and a small electrode. The resistance[12] ab is 10^5 ohms and bc is 6×10^3 ohms. A Western Electric 201A input transformer is placed between

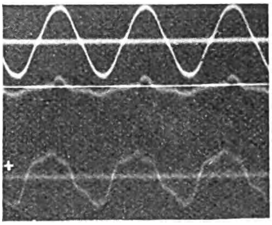

Fig. 6—Curves for 0.0004 Ampere Positive.

60~Timing wave

Rectified current

Voltage across discharge

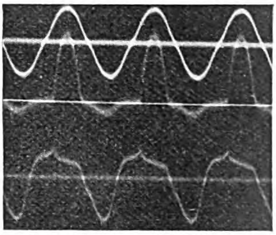

Fig. 7—Curves for 0.01 Ampere Positive.

the voltage element and the amplifier in order to eliminate from the oscillograph element the d.c. component in the plate circuit. This is necessary in order to keep the zero axis in place under normal conditions.

I. One-Way Rectification

The circuit connections for this experiment are shown in Fig. 4. The small electrode, $F+$, is shorted to the cathode, and ammeter A_1, in series with a low resistance load, is shunted across terminals A–B.

With 21.5 volts$_{eff}$[13] across the tube, a discharge, accompanied by a purplish glow,[14] appears in the region between the $F-$ electrode and the cathode. This discharge, the chief cause of luminosity of which is the negative glow, is the only discharge present in the tube at this

[12] It is important to note that with low resistances the magnitude of the reverse current is affected markedly by the value of this resistance. At the value of 10^5 ohms, the reverse current is not modified appreciably.

[13] Due to an apparent photo electron emission from the electrodes of this tube, this potential is less when the tube is exposed to the radiation from a 75-watt incandescent lamp. It is necessary therefore to exclude extraneous light when getting the potential values.

It may be mentioned that this apparently abnormally low voltage is probably due to the presence, on the electrodes, of alkali metals which operate to reduce the cathode fall. Charlton reports that the use of caesium oxide together with finely divided magnesium which is evaporated and deposited upon the inside of the tube will reduce the potential drop to one-half of what it is without the use of these. (For further data regarding the abnormal low voltage arc, cf. Holst and Osterhuis, loc. cit. Holst and Osterhuis, *Physica*, **4**, 42; 1924. Compton and Eckart, *Phys. Rev.*, pp. 139–146; Feb. 1925.)

[14] Although these glow discharges are alternating, they are seen as continuous due to persistence of vision.

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potential. It is interesting to note that Crookes' dark space about one-half millimeter in width is perceptible near the cathode during this discharge. As the a.c. potential is raised, the glow becomes brighter but is still confined in the narrow space where the small

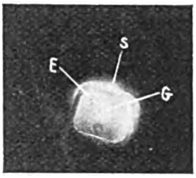

Fig. 8—Discharge at 0.01 Ampere.

electrode, $F-$, extends through the cathode. At a potential slightly in excess of the initial discharge potential, the ammeter indicates a reverse current of 0.0002 ampere. This reverse current is due to a rectifying action opposite in direction to the main rectified flow, which has not yet started.

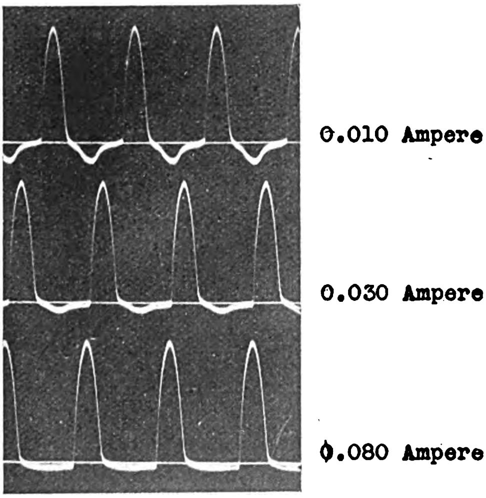

Fig. 9—Curves Showing Percentage Variation of Reverse Current with Load.

As already pointed out, this reversed rectifying action can be anticipated from the static characteristic curves given in Fig. 2. It is of interest to note further that with the tube, whose d.c. characteristics are given in Fig. 2, the negative stream or reverse current in

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both the positive and negative anode circuits starts at a much lower potential than the positive current. This enables us to anticipate the presence of reverse currents, preceding the positive currents which constitute the main rectified flow. Several tubes showed exceedingly small reverse currents and one tube required a lower striking potential for the positive current than for the reverse current. This indicates that the positive current preceded the reverse current and there was no initial reverse current. We can infer from this that the direction of asymmetry is apparently not fixed at the start thus throwing into question the ionization theory of Townsend, as already mentioned. The striking potential of all the tubes examined differed from one another as much as 25 per cent. The variation in the reverse currents is probably traceable to the differences in construction such as centering of the small electrodes in the hole in the cathode, inequalities in the deposits of alkali metals on the electrodes, etc.

Fig. 10—Typical Voltage Curve.

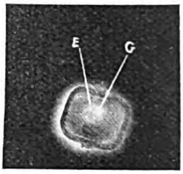

Fig. 11—Discharge at 0.05 Ampere.

If the a.c. potential is now raised to 28.5 volts$_{eff}$, the rectified stream suddenly reverses itself and the discharge from the $F-$ electrode extends over to the one side of the inner surface of the cathode. Fig. 5 shows the nature of the glow discharges. At this point, the meter indicates a positive[15] current of 0.0004 ampere. This is the initial value of the rectified current when the main discharge takes place to the cathode. When the main discharge sets in there is a sudden drop in the voltage across the discharge occurring at or near the maximum, as shown by the voltage curve[16] in Fig. 6. This continues to decrease while the positive current is building up, after which the voltage becomes negative. With a higher a.c. potential the positive current increases considerably, while the reverse current increases by only a small amount. Fig. 7 shows the curves for a rectified current of 0.01 ampere; and Fig. 8 the photograph of the corresponding glow phenomena. The reverse current does not change markedly with the

[15] In the actual experiment a middle zero instrument is best used.
[16] The sequence of events on all curves proceeds from left to right.

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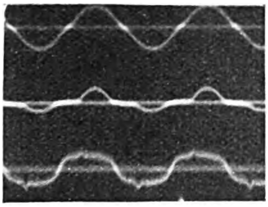
load, up to full load; nor does the curve of voltage variation across the discharge indicate any appreciable change from what it is for a load of 0.01 ampere. Current curves which show the percentage variation of the reverse current with loads of 0.01 ampere, 0.03 ampere, and 0.08 ampere, are given in Fig. 9. These curves were obtained with a variable shunt and from the practical disappearance of the



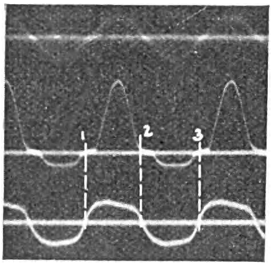


Fig. 12—Curves for 0.0004 Ampere Positive.

Fig. 13—Curves for 0.01 Ampere Positive.

reverse current, for constant positive amplitude, they show the very small variation of the reverse current[17] with load. Fig. 10 shows a typical voltage curve over a range of 0.015 to 0.08 ampere. This was taken at approximately full load. Fig. 11 shows the extent of the negative glow adjacent to the cathode for a load of 0.05 ampere.

II. Two-Way Rectification

In these experiments $F+$ was joined to the secondary of the transformer. The circuit connection is given in Fig. 4. The $F+$ electrode is disconnected from the cathode and is joined to the secondary of the transformer by means of the connection shown dotted. Under this condition the discharge is maintained alternately to the $F+$ and $F-$ electrodes. The reverse current is present but inappreciable.

When the a.c. potential reaches 18.0 $\text{volts}_{\text{eff}}$,[18] the rectifying discharge strikes with a positive current of 0.0004 ampere. Fig. 12 shows how the voltage and current vary with these currents. The voltage

[17] It is interesting to note that the reverse current can be studied in another way. Instead of shunting the increased current at higher loads, a rectifier tube with $F+$ shorted to the plate (cathode) can be connected into the circuit of Fig. 4 at X. With the cathode of this tube connected to the transformer terminal adjacent to X, the positive current is eliminated. This enables a study to be made of the reverse current for loads beyond the current capacity of the oscillograph. The effect which this may have upon the reverse current, however, makes this procedure questionable.

[18] This is one-half of secondary voltage as in preceding case.

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wave shows a double break which is due to the discharge alternating between the $F+$ electrode and the $F-$ electrode. The current wave is taken in the $F-$ electrode circuit and shows a positive current slightly greater than the reverse current. When the current wave was taken in the ammeter circuit, it possessed two positive lobes, each one identified with the positive and negative breaks in the voltage wave. As the potential is raised to 66.0 volts$_{eff}$, the sharp breaks in the voltage wave

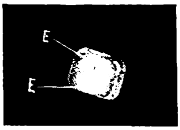

Fig. 14—Discharge at 0.01 Ampere.

disappear. Fig. 13 shows the curves for a current of 0.01 ampere at 66.0 volts$_{eff}$. The photograph of the discharge at this current is given in Fig. 14, which shows a glow at each of the small electrodes.

It is of interest to point out in Fig. 13 that at point 1, where the large electrode has become negative, there is a delay in the striking of the rectifying discharge until the potential has built up to a certain value. At point 2 it is seen that there is a tendency for the current to maintain itself even after the voltage has reversed. This post-arc

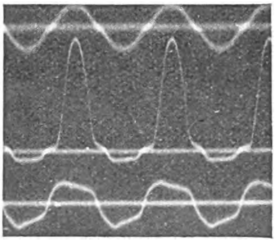

Fig. 15—Curves for 0.02 Ampere Positive.

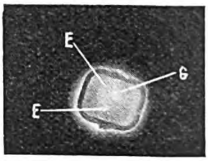

Fig. 16—Discharge at 0.05 Ampere.

conductivity[19] is probably due to a diffusion of positive ions into the large electrode even after the reversal of the voltage. At point 3 the reverse current ceases as soon as the large electrode becomes negative again, at the end of the cycle.

At 90.0 volts$_{eff}$ a current of 0.01 ampere is available. The current and voltage waves are given in Fig. 15. There is very little difference

[19] M. L. Pool, *Phys. Rev.*, 30, 848; 1927.

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between the current and voltage waves at 0.01 ampere, and 0.02 ampere, as shown in Figs. 13 and 15. Fig. 16 shows the nature of the glow discharges at 0.05 ampere. With this current the glow extends outside the gauze cover, as would be expected from the theory of the "normal" cathode currents and fall of potential.

III. Two-Way Rectification with Filter Circuit

In these experiments the effect of the filter circuit on the reverse current is studied. The circuit connections are given in Fig. 4. The $F+$ and $F-$ electrodes are connected to the transformer. Ammeter A_1 with the low resistance load is disconnected from $A-B$ and the filter circuit, shown at the right, is connected to terminals $A-B$. This is a standard low-pass filter circuit embracing two condensers of about 2 μf capacity each, and an inductance of 20–30 henries. The load resistance is 3300 ohms.

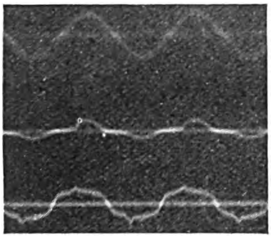

Fig. 17—Curves for 0.0009 Ampere Positive.

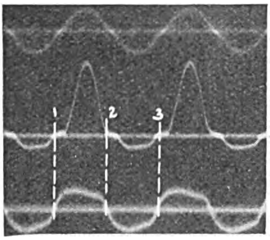

Fig. 18—Curves for 0.01 Ampere Positive and 26 Volts Direct Current.

The initial rectified discharge current appears suddenly and has a value of 0.0009 ampere, as measured at A_2. Before this there is a very small reverse current. The voltage and current waves are shown in Fig. 17. These waves are the same as those shown in Fig. 12, taken without the filter in the circuit. The effect of the filter is more clearly shown at higher current values.

As the a.c. potential is raised, the sharp break in each lobe of the voltage wave is less obvious, until at 68.0 volts$_{eff}$ and a load of 0.01 ampere, it is hardly visible. Under these conditions, the d.c. potential across the load is 26.0 volts. When the a.c. potential is 115.0 volts$_{eff}$ the load current is 0.02 ampere, and the d.c. potential is 50.0 volts. The voltage and current waves under these two conditions are given in Figs. 18 and 19. The discussion of points 1, 2, and 3 in Figs. 18 and 19 is similar to that given in connection with Fig. 13.

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In the cases of one-way and two-way rectification without the filter, the voltage across the electrodes in the gas was the a.c. potential applied. With the filter in the circuit the d.c. potential across the load is added to the a.c. potential. The effect of this is to increase the

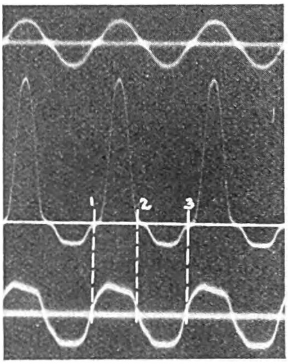

Fig. 19—Curves for 0.02 Ampere Positive and 50 Volts Direct Current.

magnitude of the reverse current as the d.c. voltage increases. This increase in the reverse current is clearly seen in Figs. 18 and 19, for the case of the filter, as compared with the earlier curves in Figs. 13 and 15. Further work remains to be done, using various filter combinations, in order to determine how the reverse current can be minimized, since it is the magnitude of the reverse current that fixes the rectifying characteristics of this type of tube.

Performance as Rectifier

The relative magnitudes of the positive current and the reverse current in a rectified wave are an indication of the degree of asymmetry. In one-way asymmetry, if the reverse current is entirely suppressed, half-wave rectification is complete. In two-way asymmetry, if there

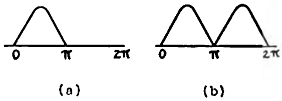

Fig. 20—Theoretical Half- and Full-Wave Rectified Cycle.

is a conversion, so to speak, of the negative lobe of the cycle into a full positive lobe without any reverse current, full-wave rectification is complete. These two conditions are shown in Fig. 20. In Fig. 20a, the interval $0 \rightarrow \pi$ represents the time during which the large electrode

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is functioning as cathode and the greatest current flow takes place. The interval $\pi \rightarrow 2\pi$ represents the time during which the large electrode is functioning as anode and is suppressing the reverse current. In Fig. 20b, there are two positive lobes resulting from the discharge alternating in only one direction between the cathode and either of the small electrodes.

The degree of asymmetry or degree of rectification is sometimes expressed as a percentage, although this does not clearly represent its meaning. Since the degree of rectification refers more exactly to the proportion of alternating current converted into direct current, Holler and Schrodt[20] prefer to define it as the ratio of the average value of the current, as read on a permanent magnet type of meter, to the root mean square value, as read on a dynamometer or thermal type instrument. In general,

$$\text{Degree of Rectification} = \frac{\text{d.c.}}{\text{a.c.}} = \frac{\text{Average value of current}}{\text{r.m.s. value of current}}$$

In half-wave rectification, for complete suppression of the negative lobe, we have, applying the mean value theorem to Fig. 20a,

$$\frac{\text{d.c.}}{\text{a.c.}} = \frac{\displaystyle\int_0^\pi \frac{I_m \sin\xi \, d\xi}{2\pi}}{\sqrt{\displaystyle\int_0^\pi \frac{I_m^2 \sin^2\xi \, d\xi}{2\pi}}} = \frac{2}{\pi} = 0.636$$

The value of the preceding ratio, 0.636, represents the maximum degree of half-wave rectification possible, when the circuit contains pure resistance and no counter electromotive force. Similarly for full-wave rectification which is theoretically complete we have,

$$\frac{\text{d.c.}}{\text{a.c.}} = \frac{2\displaystyle\int_0^\pi \frac{I_m \sin\xi \, d\xi}{2\pi}}{\sqrt{2\displaystyle\int_0^\pi \frac{I_m^2 \sin^2\xi \, d\xi}{2\pi}}} = \frac{2\sqrt{2}}{\pi} = 0.905$$

The value of this ratio, namely 0.905, represents the maximum degree of full-wave rectification possible under the above conditions.

In order to determine these two ratios, a permanent magnet type of ammeter and a thermogalvanometer are connected in series in

[20] Technologic Papers of Bureau of Stds. No. 265, Oct. 1924, p. 472.

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place of ammeter A_1 in Fig. 4, the filter circuit being removed. With this arrangement the following data were collected:

	Half Wave a.c.ma = 100.0		Full Wave a.c.ma = 100.0	
Tube	$d.c._{ma}$	$\frac{d.c.}{a.c.}$	$d.c._{ma}$	$\frac{d.c.}{a.c.}$
1	50	0.50	78	0.78
2	61	0.61	84	0.84
3	58	0.58	78	0.78
4	59	0.59	80	0.80
5	60	0.60	84	0.84

Jolley defines a so-called rectification ratio as the ratio of the mean current in one direction to that in the other. This ratio is proportional, approximately, to the ratio of the surface areas of the large electrode and the small electrode. Hence the degree of asymmetry is proportional to these surface areas. In the two-electrode tube investigated by Kneser this ratio is 380/1, whereas in Charlton's and the one described in this paper the ratio is approximately 300/1.

The preceding table of data shows that the ratio of the surface areas does affect the degree of rectification since for tube 1, with the gauze cover and hence a smaller cathode area, the ratio of d.c./a.c. is noticeably smaller.

The author acknowledges his great indebtedness to Prof. Dempster, under whose direction this work was carried out, for the suggestion of the problem and for his help in the interpretation of the results. The author is also indebted to Dr. Hoag for his assistance, in many ways, during the course of the work.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

VACUUM-TUBE VOLTMETER DESIGN*

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Summary—*It is shown that for every input range and meter a set of optimum conditions exist, which insure maximum readability, economy, and minimum operating current. Equations are developed (and illustrated with several examples) for the rapid design of a meter for a predetermined range. Optimum conditions are imposed.*

The dependence of the range secured upon plate circuit impedance is shown and its variation with frequency explained. The procedure for eliminating frequency error up to and including radio frequencies is illustrated.

Perhaps no instrument has contributed more toward making the scientific design of radio receivers a practical reality than has the vacuum-tube voltmeter. It fills a position in the communications field of as great an importance as the current-operated voltmeter holds in power work. It has supplied a measuring rod where none before existed. It represents one of man's nearest approaches to the ideal voltmeter.

But despite its excellent characteristics and its extended usefulness, little has been done to make its design a straightforward process. All too often vacuum-tube voltmeter design, if it may be called such, has consisted of calibrating a heterogeneous assembly of parts with little regard for the range secured or the maximum performance obtainable from the setup. Jansky and Feldman[1] have made about the only advance out of the forest of cut-and-try methods with their equation for the maximum peak voltage measurable in a continuous two-range design.

For every desired range and tube a set of optimum circuit conditions exist, a fact not generally appreciated. It is possible to predetermine the circuit constants to give any desired range without extensive experimenting to find them. It is the purpose of this paper to consider optimum conditions in deriving certain equations, and to consider the effect of frequency on the results.

The Optimum Concept

Fig. 1 shows a typical vacuum-tube voltmeter circuit and Fig. 2 a family of curves giving its performance under various operating

* Dewey decimal classification: R 261. Original manuscript received by the Institute, January 18, 1929.

[1] Jansky and Feldman, "A Two-Range Vacuum Tube Voltmeter." *Jour* A.I.E.E., **XLII**, 126; February, 1929.

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conditions. These curves represent the results of a considerable number of cut-and-try tests undertaken in the design of the low reading 199-tube meter shown in Fig. 4. That the range increases with the value of the plate circuit series resistor, R_p', is evident; due, of course, to the greater increase in plate voltage drop across the resistor with current as the R_p' value increases. The circles on curves *C* and *F* show

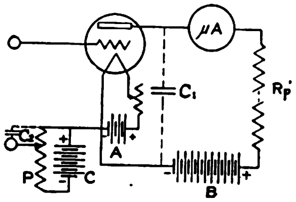

Fig. 1—Wiring Diagram of Vacuum-Tube Voltmeter.

the start of grid-current flow and indicate insufficient grid bias. Reference to the table of voltages accompanying Fig. 2 shows that an excess bias was used for curve *D*, requiring an abnormally large plate voltage to counteract it, and resulting in the plate meter reaching its maximum deflection before the limit of grid bias was approached. For certain other curves, as *A*, *B*, or *E*, the constants were more pro-

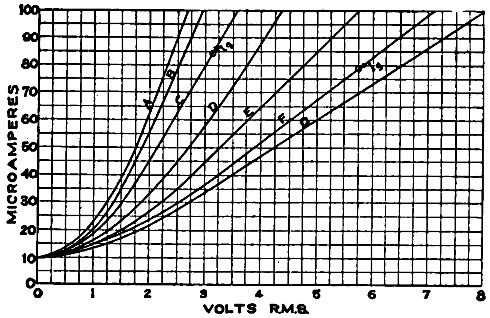

Fig. 2—Variation of Vacuum-Tube Range with Circuit Constants.

perly chosen, the grid bias being approached but not exceeded by the input wave, the plate voltage being normal, and the limit of plate meter deflection reached simultaneously with that of the grid bias.

TABLE OF VOLTAGES

Curve	Range, E	E_g	E_p	R_p'
A	2.7 v	−4.4 v	23 v	10,000 ohms
B	3.0	−4.6	23	25,000
C	3.6	−4.5	23	75,000
D	4.4	−9.0	48	100,000
E	5.7	−9.0	48	200,000
F	7.0	−8.7	48	300,000
G	8.0	−13.0	71	350,000

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By reasoning further, the *optimum conditions* are seen to be that set of circuit constants wherein the desired range is spread over the whole indicating meter scale, wherein the minimum plate and grid potent tials are employed, and wherein *the grid of the tube at no time becomes positive.* Maximum readability, economy, and true voltmeter action are thus assured. The importance of keeping the grid always negative is often underrated, but the effect of an appreciable operating current on the circuits used in communication work is altogether too potent to be neglected.

Design Equations

Starting with the theorem that the grid must always remain negative and recalling the fact that the peak value of a sine wave is 1.41 times its root-mean-square value we can write:

$$E_g = 1.5E \tag{1}$$

where E_g is the grid bias and E is the maximum r. m. s. voltage to be measured. The constant 1.5 is selected to allow in some measure for wave forms more sharply peaked than the sine wave.

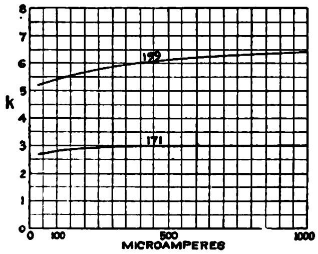

Fig. 3—Values of k for Various Maximum Plate Currents.

From the relation for the plate current of the tube, an expression for the value of the plate circuit series resistor can be secured, i.e., from

$$I_p = \frac{\mu E}{R_p + R_p'} \tag{2}$$

where I_p is the plate current, μ the amplification factor of the tube, R_p the tube plate impedance, and R_p' the plate-circuit series resistance.

Now I_p is the maximum reading of the plate meter minus the steady false zero reading for full scale deflection of the meter

$$I_p = Im - Io = I_p'. \tag{3}$$

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Also, let

$$\mu = k \tag{4}$$

where k is the value of the amplification constant for very large values of grid bias.

Substituting (3) and (4) in (2) gives

$$I_p' = \frac{kE}{R_p + R_p'}$$

and

$$I_p'R_p + I_p'R_p' = kE$$

or

$$R_p = \frac{kE - I_p'R_p}{I_p'}. \tag{5}$$

The plate voltage required, E_p, is given quite closely by

$$E_p = kE_g. \tag{6}$$

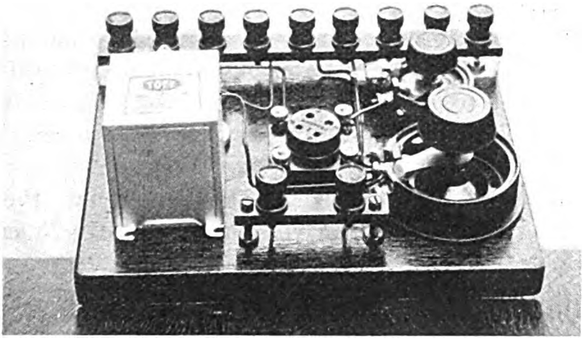

Fig. 4—Low-Range 199 Vacuum-Tube Voltmeter.

Values of k for 199 tubes and 171 tubes are given in Fig. 3 for various values of microammeter maximum range.

Experimental Verification

The validity of the foregoing equations can be established by a consideration of several examples. Take first the A range of Fig. 2 (which is later used in the analysis of the effect of frequency). A 199 tube and 0-100-μa plate meter are used to give an input voltage range of 0-2.7 v.

From (1)

$$E_g = 1.5 \times 2.7$$
$$= 4.05 \text{ v}$$

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For the 199 tube and 0-100 microammeter $k=5.4$ from Fig. 3. Also $I_p'=100-10$ μa$=0.00009$a. $R_p=150{,}000$ ohms. Therefore, in (5):

$$R_p'=\frac{5.4\times 2.7-0.00009\times 150{,}000}{0.00009}$$

$$=12{,}000 \text{ ohms.}$$

From (6)

$$E_p=5.4\times 4.05=21.8 \text{ v}$$

The values used, as given for curve A in the table accompanying Fig. 2, were: $E_g=4.4$ v, $E_p=23$ v and $R_p'=10{,}300$ ohms, a larger plate voltage supplied by a new 22 1/2-v block requiring a correspondingly larger grid bias.

Calculations for curve E, a range of 5.7 volts, give: $E_g=8.7$ v $E_p=47$ v, and $R_p'=197{,}000$ ohms, against $E_g=9.0$ v, $E_p=48$ v, and $R_p'=200{,}000$ ohms as used.

A valuable independent check is secured by calculating the constants for a voltmeter designed by E. T. Dickey and published in 1927.[2] A 171 tube and a plate meter of 0-100 μa with grid biases of 9 and 90 v were used to give voltage ranges of 0-7 v and 0-70 v, respectively. Under the criteria of optimum conditions, however, these ranges are more nearly 0-6 v and 0-60 v, since the grid swings positive for values of 6.4 v and 64 v with a sine wave shape. Since the maximum value of the plate meter is not reached, $I_p'=85-10=75$ μa, and $=80-10=70$ μa.

For the 60-volt range

$$E_g=1.50\times 60=90 \text{ v} \qquad (1)$$

$$R_p'=\frac{2.8\times 60-0.00007\times 10{,}000}{0.00007}$$

$$=2{,}390{,}000 \text{ ohms} \qquad (5)$$

$$E_p=2.8\times 90=250 \text{ v} \qquad (6)$$

The values given are; $E_g=90$ v, $E_p=250$ v, and $R_p'=2{,}400{,}000$ ohms. Calculations for the 0-6 v range give: $E_g=9$v, $E_p=25$ v, and $R_p'=214{,}000$ ohms, as against $E_g=9$ v (max.), $E_p=22.5$ v, and $R_p'=200{,}000$ ohms.

[2] E. T. Dickey, "Notes on the Testing of Audio Frequency Amplifiers," Proc. I.R.E., 15, 687; August, 1927.

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The agreement between calculated and actual values is thus seen to be very good, a slight change in grid bias effected by adjusting the potentiometer to give the correct zero reading generally serving to make it exact. The simplicity of the calculations is apparent.

It is interesting to note that the plate impedance at which a tube is worked is very closely a constant for a given plate meter over the range of the plate and grid potentials used. For a 199 tube and 0-100-μa meter R_p is 150,000 ohms, very nearly, for values of negative grid bias from 4.4 to 13 v, corresponding to plate voltages from 23 to 71, and plate circuit series resistors from 10,000 to 350,000 ohms. For a

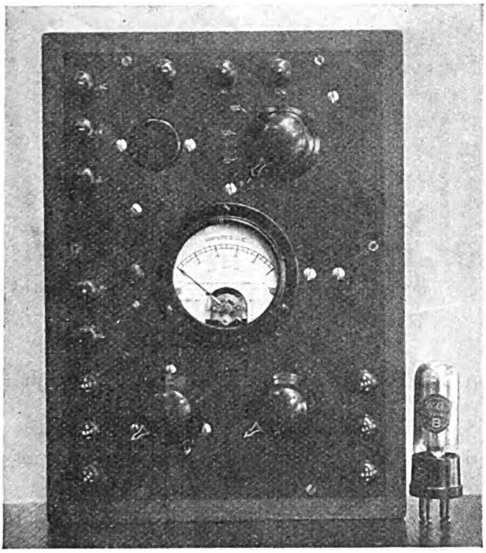

Fig. 5—Panel-Type 199 Vacuum-Tube Voltmeter.

199 tube and 1,000-μa meter R_p is 50,000 ohms, and for a 171 tube and 100-μa meter, 10,000 ohms. These are the values existent for full scale deflection.

Effect of Frequency

The design equations presented give the range secured for a plate circuit path of relatively low impedance, that resulting from resistance alone. Should frequency make the effect of inductive or capacitive reactance appreciable the range secured is noticeably affected. Fig. 6 shows the performance of the voltmeter shown in Fig. 4 at various frequencies without plate circuit bypassing.

It will be noted that changes in frequency from zero to radio fre-

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quencies cause variations in the voltage range secured of more than 25 per cent each side of the mean. Omitting for the moment the direct-current calibration, which never can coincide with any alternating-current range, three zones of calibration are observed; one secured with a low-frequency alternating current, the 60-cycle curve; a second secured at higher audio frequencies, the 500-, 5,000-, and 16,000-cycle curves; and a third secured at radio frequencies, the 1,000,000-cycle curve.

The variation of range with frequency is seen to be non-linear, increasing frequency first increasing the range and later decreasing it. This can be explained by considering the plate-circuit meter. A mi-

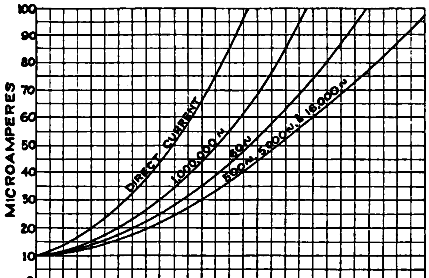

Fig. 6—Variation of Range with Frequency Plate Circuit *Not* Bypassed by Condenser C_1.

croammeter moving coil possesses considerable self inductance because of its many turns and close proximity to considerable iron, and it also possesses an appreciable distributed capacitance as a result of the concentrated nature of the winding and the shellac varnish used to impregnate it. At 60 cycles the effect of capacitance is negligible, but that of inductance has some effect on determining the impedance. At the higher audio frequencies the inductive reactance becomes considerable, and the plate current secured for a given input decreases, as is indicated by the lower position of this curve. This effect continues for higher frequencies, augmented by the parallel resonance effect of the inductance and distributed capacitance, until the resonant frequency of the combination is reached. Beyond this point the coil acts as a condenser and by the time radio-frequency currents are reached it offers little impedance to their flow. Similar effects in the resistor, batteries, and wiring all result in making the impedance a minimum and thus place the radio-frequency curve above the others.

Referring now to Fig. 7, it is seen that with the plate circuit bypassed by C_1 ($4\mu f$) and an incidental condenser C_2 ($1\mu f$) across the

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potentiometer of Fig. 1, the range secured is independent of frequency. The various points around the curve represent the plotting of data for 60, 500, 16,000, and 1,000,000 cycles, all of which are so nearly coincident that only a single curve can be drawn through them. The impedance of the plate circuit has been made a constant regardless of frequency by the use of the condenser.

The d.c. calibration differs from all a.c. curves for two major reasons. First, the value of the a.c. input wave read is the *root-mean-square* value as indicated by a.c. meters, whereas the plate-current meter, being a d.c. instrument, gives the *average* value of the plate-current wave, or the integrated semi-constant flow if a bypass condenser be used. The "average" value of a d.c. voltage is indicated by

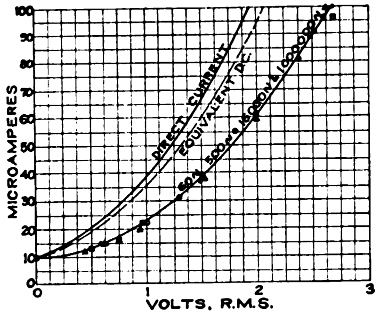

Fig. 7—Variation of Range with Frequency Plate Circuit Bypassed by Condenser C_1.

a d.c. voltmeter, consequently the d.c. voltage impressed is more effective in producing plate meter deflection by the ratio of r.m.s. to average values, or by the *form factor* of the a.c. wave. The "equivalent d.c." curve of Fig. 7 gives this correction. Secondly, the a.c. plate-current waves contain appreciable "negative" loops as a result of incomplete rectification which lowers its average value considerably, i. e., "negative" values below the 10 μa false zero reading. The d.c. input, on the other hand, does not produce any "negative" loops in the plate current but merely a steady d.c. value, which fact accounts for the discrepancy between the equivalent d.c. curve and the a.c. curves of Fig. 7.

Conclusions

Of the various combinations of circuit conditions possible for the attainment of a certain voltage range, there exists *one* set of optimum conditions. This set of conditions spread the desired range over the

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entire scale of the indicating meter, require the minimum plate and grid battery potentials, and at no time allow the grid of the tube to become positive. Through the use of the design equations developed above, the optimum conditions that will give any desired range are secured at once.

Variation of range with frequency is chiefly due to variation of plate-circuit impedance, a quantity affected mainly by the properties of the plate-circuit meter. Frequency error can be reduced to a negligible quantity by adequately bypassing the plate circuit with a "tank" condenser. Direct-current calibration curves cannot coincide with those for alternating current, because of an inherent discrepancy in the instrument indications amounting to the value of the form factor, and because of "incomplete" rectification, the latter factor usually being the greater of the two.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

FIELD INTENSITY CHARACTERISTICS OF DOUBLE MODULATION TYPE OF DIRECTIVE RADIO BEACON*

By
HARADEN PRATT

(Mackay Radio and Telegraph Company, New York City)

Summary—*Field strength characteristic relationships, applicable to the double modulation type of directive radio beacon when modulation is applied, are developed. Graphs showing the characteristic patterns which govern the indications of the receiving apparatus as it moves around in space about the transmitter are given for three typical sets of transmitter adjustments.*

THE publication in the PROCEEDINGS[1] of details concerning the double modulation type of directive radio beacon as developed at the Bureau of Standards did not discuss the relationships necessary to obtain field intensity characteristic patterns when modulation is applied to each of the carrier frequency currents of the system.

It is necessary to know what such relations are in order to predict accurately what deflections the indicating device will exhibit at the receiving station for different positions in space around the beacon transmitter.

The writer has worked out these relations and they are presented here to complete the analysis as first set forth in the previous paper.

Reference is made to (4) on page 908,[1] which states that the field intensity observed at a fixed point P is,

$$E_p = AK[E_{s_1} \cos(\theta - \delta) - E_{s_2} \sin(\theta - \delta)] \tag{1}$$

where E_{s_1} and E_{s_2} are the voltages supplied by the amplifiers to the two input coils of the antenna system goniometer, respectively.

When modulation is introduced, these two voltages can, for the general case, be written:

$$E_{s_1} = (E_0 + d_1 E_0 \sin \omega_1 t) \sin \omega t \tag{2}$$

$$E_{s_2} = r(E_0 + d_2 E_0 \sin \omega_2 t) \sin(\omega t + \beta) \tag{3}$$

Where E_0 is the maximum value of the carrier voltage supplied by the first amplifier.

r is the ratio of the maximum value of the carrier supplied by the second amplifier to that supplied by the first amplifier.

* Dewey decimal classification: R 526.1. Original manuscript received by the Institute, January 21, 1929.

[1] J. H. Dellinger and Haraden Pratt, Proc. I. R. E., 16, 890; July, 1928.

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d_1 is the ratio of the maximum value of the voltage of modulation frequency $\omega_1/2\pi$ to E_o.

d_2 is the ratio of the maximum value of the voltage of modulation frequency $\omega_2/2\pi$ to rE_o.

β is the phase angle between the two carrier voltages.

These values when placed in (1) result in

$$
\begin{aligned}
E_p = AKE_0 \Big\{ & \sin \omega t [\cos(\theta-\delta) - r \sin(\theta-\delta)\cos\beta] \\
& - r \cos \omega t \sin\beta \sin(\theta-\delta) \\
& + \frac{d_1}{2}[\cos(\omega-\omega_1)t - \cos(\omega+\omega_1)t] \cos(\theta-\delta) \\
& - \frac{rd_2}{2}[\cos(\omega t+\beta-\omega_2 t) - \cos(\omega t+\beta+\omega_2 t)] \sin(\theta-\delta) \Big\}
\end{aligned}
\tag{4}
$$

which is the general expression for the field intensity at the receiving point.

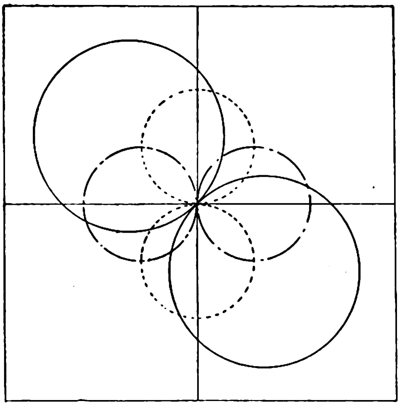

Fig. 1—Field Intensity Patterns Showing Carrier and Both Sets of Sidebands of Double Modulation Beacon for the Case Where the Two Carrier Voltages Supplied to the Antenna Systems Are Equal and in Phase.

Several special cases of this general relationship are of interest, particularly those representing possible operating conditions. One usual operating condition is offered by the case where the two carrier voltages have the same amplitude and are in phase. Then $r=1$, $\beta=0$, and

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$$E_p = AKE_0 \left\{ \sqrt{2} \sin \omega t \sin\left(\; -\delta - \frac{\pi}{4} \right) \right.$$

$$+ \frac{d_1}{2} [\cos (\omega - \omega_1)t - \cos (\omega + \omega_1)t] \cos (\theta - \delta) \tag{5}$$

$$\left. - \frac{d_2}{2} [\cos (\omega - \omega_2)t - \cos (\omega + \omega_2)t] \sin (\theta - \delta \right\}$$

Fig. 1 shows the field intensity patterns for this case, the radius vector being proportional to the field intensity for any angle in space referred to the radio-beacon transmitter. The solid line graph represents that of the carrier which is indicated by the first term of the equation, and the dot-dash and dotted graphs those of the sidebands resulting from the two modulating frequencies respectively, indicated by the two other terms of the equation.

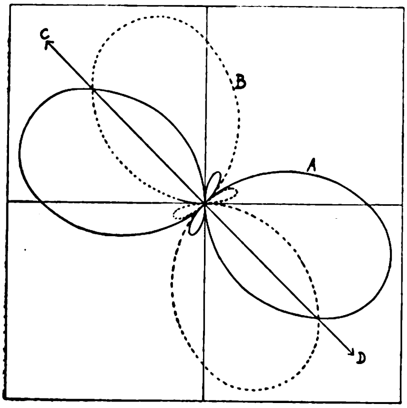

Fig. 2—Field Intensity Patterns Showing Reed Indicator Deflections for the Conditions Shown in Fig. 1.

Since the reed course indicators developed for use with this type of radio-beacon signal operate as a result of the low-frequency voltages produced by the beating in the radio receiver of the sideband frequencies upon that of the carrier, they will respond to space pattern characteristics as shown in Fig. 2, which is easily derived from Fig. 1. The solid graph corresponds to the voltage available for the operation of one reed, and the dotted graph that for the other reed. No corrections have been allowed to care for any non-linearity which the radio receiver circuits may introduce, for example by a non-linear detector. The courses are shown at C and D. Practically no deflections occur in the directions at right angles to these courses. The sharpness of the course indications can be estimated from the patterns.

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This operating condition appears to be the most favorable so far investigated. The simplification resulting from only two courses being produced is of great advantage.

The case where the two carrier frequency voltages are unequal and in phase is of interest. Then $r<1$, $\beta=0$, and

$$E_p=AKE_0\Big\{[\sin \omega t[\cos(\theta-\delta)-r\sin(\theta-\delta)]$$
$$+\frac{d_1}{2}[\cos(\omega-\omega_1)t-\cos(\omega+\omega_1)t]\cos(\theta-\delta)$$
$$-\frac{rd_2}{2}[\cos(\omega-\omega_2)t-\cos(\omega+\omega_2)t]\sin(\theta-\delta)\Big\}. \quad (5)$$

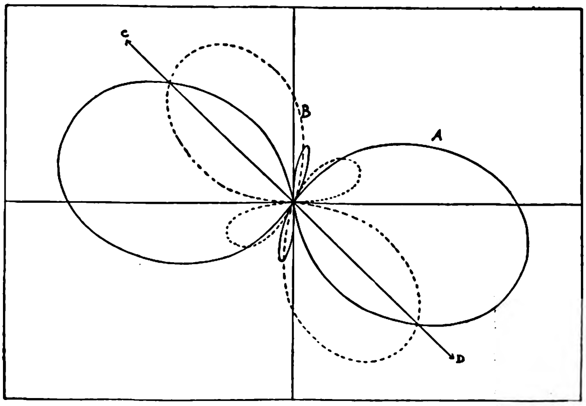

Fig. 3—Field Intensity Patterns Showing Reed Indicator Deflections for the Case Where the Two Carrier Voltages Supplied to the Antenna Systems Are Unequal but in Phase.

Fig. 3 shows the patterns for reed deflections for a typical set of conditions coming under this case. It will be noted that this results in an unsatisfactory operating condition, the courses not being well defined from the standpoint of reed deflections.

By suitable coupling means from the master oscillator to the two output circuits, one of the carrier voltages can be advanced in phase beyond that of the other. The case where the two carrier voltages are equal in amplitude but in phase quadrature results in a different field intensity pattern, the resultant radio carrier voltage vector rotating in space so that at all points equidistant from the transmitter, the field intensity changes periodically in time at radio frequency, but at the same amplitude. Here $r=1$, $\beta=\pi/2$, and,

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$$E_p = AKE_0 \Big\{ [\sin \omega t \cos (\theta - \delta) - \cos \omega t \sin (\theta - \delta)]$$
$$+ \frac{d_1}{2} [\cos (\omega - \omega_1)t - \cos (\omega + \omega_1)t] \cos (\theta - \delta)$$
$$- \frac{d_2}{2} [\sin (\omega + \omega_2)t - \sin (\omega - \omega_2)t] \sin (\theta - \delta) \Big\}. \quad (7)$$

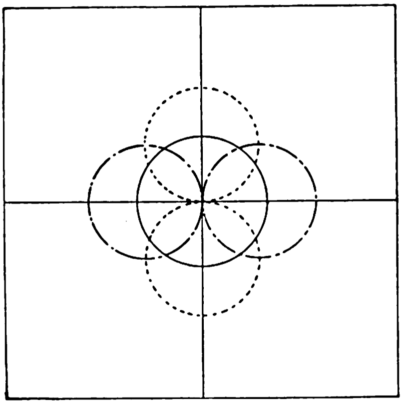

Fig. 4—Field Intensity Patterns Showing Carrier and Both Sets of Sidebands of Double Modulation Beacon for the Case where the Two Carrier Voltages Supplied to the Antenna Systems Are Equal but in Quadrature Phase.

The space pattern for this case is shown in Fig. 4, and the variation of the audio-frequency intensities producing the reed deflections derived from it is shown in Fig. 5. Four courses result. It is interesting

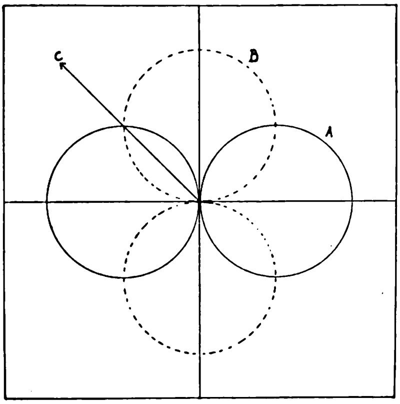

Fig. 5—Field Intensity Patterns Showing Reed Indicator Deflections for the Conditions Shown in Fig. 4.

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to note that since the carrier is of equal strength at all points, this arrangement might also permit the use of the beacon for some other purpose, for example with direction finders installed on aircraft, using the heterodyne method of reception.

The relations given for the last case are, of course, subject to the variations discussed in (17) and (18) of the original paper for which allowance must be made.

Relations between the two carrier currents other than those discussed can be readily studied through examination of the resulting field intensity patterns.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

LOGARITHMIC SCALE FOR BEAT-FREQUENCY OSCILLATOR*

By
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Summary—*The equation governing the shape of a condenser's plates, which if placed in a beat-frequency oscillator would give a logarithmic frequency scale, is developed.*

BECAUSE of the ease with which their frequency can be varied and because they make it continuously variable, beat-frequenyc oscillators are rapidly gaining favor. A change in frequency is accomplished by simply changing the capacity setting of one variable air condenser. Its scale can be calibrated directly in frequency. This saves the operator a great deal of time and the chances of error are greatly reduced.

It has become general practice to plot audio-frequency curves and sound pressure curves on semi-log coordinate paper. This is in order that the low end of the frequency scale will be spread out. By doing this the low frequencies can be plotted to the same number of significant figures as a frequency ten or a hundred times higher. Likewise in setting the frequency of a beat-frequency oscillator it is very desirable to be able to set say ninety cycles to the same number of significant figures as nine hundred or nine thousand. This will be possible if the frequency scale is a logarithmic one.

Let us consider the shape of a condenser plate which will give a logarithmic frequency scale. The condition to be satisfied is

$$\frac{d(\log f)}{d\theta}=K \tag{1}$$

Where θ = the angle through which the condenser is rotated
K = a constant
and f is the audio frequency

Now $f=f_1-f_2$ (2)

where f_1 equals a constant frequency and f_2 is a variable frequency. The frequency f_2 will depend upon the circuit constants.

$$f_2=\frac{1}{2\pi\sqrt{LC}}\cdot \tag{3}$$

* Dewey decimal classification: R344.8. Original manuscript received by the Institute, December 6, 1928.

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As C will be the only variable we can write

$$f_2=\frac{D}{\sqrt{C}} \qquad \text{where} \qquad D=\frac{1}{2\pi\sqrt{L}}\,. \tag{4}$$

Substitute (4) in (2)

$$=f_1-\frac{D}{\sqrt{C}}\,. \tag{5}$$

Substitute (5) in (1) and integrate

$$\log\left(f_1-\frac{D}{\sqrt{C}}\right)=K\theta+k. \tag{6}$$

The constant of integration k can be determined from the relation
When $\theta=0$
$C=C_0$

C_0 equals the effective capacity in the circuit when θ equals zero.

Hence $(f_1-D/\sqrt{C_0})=f_0=$ the lowest audio frequency the oscillator is designed to produce.

Thus
$$k=\log\left(f_1-\frac{D}{\sqrt{C_0}}\right)=\log f_0. \tag{7}$$

Solve (6) for C

$$f_1-\frac{D}{\sqrt{C}}=\epsilon^{K\theta+k} \tag{8}$$

or

$$C=\frac{D^2}{[f_1-\epsilon^{K\theta+k}]^2}\,. \tag{9}$$

Let us now turn our attention to the capacity C. It is composed of the capacity of the variable condenser plus the rest of the effective capacity in the circuit which has been defined above as C_0
Let $C_c=$ Capacity of variable condenser.

Then $C=C_c+C_0$. (10)

Now the capacity of a flat plate condenser is proportional to its area

$$C_c=BA \tag{11}$$

neglecting edge effects, where A is the area of the plates and B is the proportionality factor.

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Where the movable plates are attached to a shaft the active plate area is: (See Fig. 1)

$$A = \frac{1}{2} \int [r^2 - r_1^2] d\theta. \tag{12}$$

Substituting (11) and (12) in (10) gives

$$C = C_0 + \frac{1}{2} B \int [r^2 - r_1^2] d\theta. \tag{13}$$

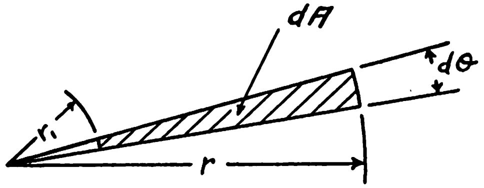

Fig. 1

Differentiate (13) with respect to θ

$$\frac{dC}{d\theta} = \frac{B}{2} [r^2 - r_1^2]. \tag{14}$$

Differentiate (9) with respect to θ

$$\frac{dC}{d\theta} = \frac{2KD^2 \epsilon^{K\theta+k}}{[f_1 - \epsilon^{K\theta+k}]^3} \tag{15}$$

We can now equate (14) and (15) and solve for r

$$r = \sqrt{\frac{4KD^2 \epsilon^{K\theta+k}}{B [f_1 - \epsilon^{K\theta+k}]^3} + r_1^2}. \tag{16}$$

Thus the radius of the plate for any value of θ is determined.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

RADIO INTERFERENCE*

By
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Summary—*The radio interference situation is rapidly becoming so serious as to make improvements in both methods of detection and methods of elimination absolutely imperative.*

The most effective method in interference detection is that combining the good features of the intensity method, with the circuit selection features of the directional method, coupled with extensive experience, both in the field and in the laboratory.

A light, self-contained, compact, unidirectional receiver equipped with a visible indicating audibility meter is the proposed instrument most desired for this work.

The design of filter systems intended to drain interference from circuits should incorporate suitable impedances at points where reflection is likely to occur.

RADIO interference has been with us ever since the inception of radio, but it is only recently that it has been receiving intensive study by the majority of central station engineers.

The first receiver, having few tubes and inefficient design, would not pick up noises from any great distances. Operators were hindered at times by interference, but the broadcast listeners regarded the radio noises in a very tolerant manner on account of the novelty of the entire radio idea. The novelty rapidly wore off and demands were made for the elimination of all noises. This demand has built up in proportion to the number of receiving sets in use, times average set sensitivity, and times a certain factor representing the proportion increase in the use of electrical appliances capable of producing oscillating discharges or surge effects.

The problem of radio interference has grown at such a rate that adequate consideration and development work have been impossible. Radio reception has become one of the greatest sources of enjoyment in life for the average person, and as an inevitable result higher standards of clear reception will soon be demanded.

Radio interference is a general term applied to anything which prevents perfect radio reception.

It is primarily applicable to noises from external sources, but it is also applied by some writers to other factors which interfere with receptions, such as fading, noises produced by receiver defects, and noises produced in eliminators and run down batteries, loose ground clamps, or bad antenna contacts.

* Dewey decimal classification: R170. Original manuscript received by the Institute, February 15, 1929. Presented before meeting of Pittsburgh section, January 15, 1929.

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Of the noises from external sources there are several classes:

(a) Static or atmospheric
(b) Station heterodyning
(c) Regenerative squeals
(d) Sparking apparatus

The work of interference detection and elimination is concentrated on the latter class. There is a great difference between an arc or uninterrupted current flow as used in welding, arc light circuits, and electric furnaces and the interrupted arc-overs across insulating mediums encountered in nearly all noise producing apparatus. The first group

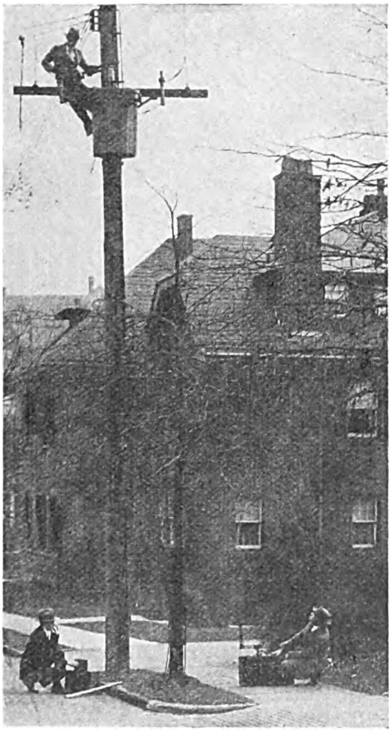

Fig. 1

uses a steady electronic flow with practically no tendency toward oscillations and few surge effects. The latter group has interrupted electronic movements productive of damped or sustained oscillations or of surge effects, the character of these depending on the constants of

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the connected circuits, the strength of currents broken, and the gap distance. The constants of the circuits include all nearby capacities, the high-frequency surge impedance of each line, adjacent inductances, and the included resistances. These are found in practice to be of such complex nature that only in the simplest cases can the results be predicted.

It may help one to visualize a minute spark as producing a terrific racket in your radio by thinking of it as a small spark transmitter similar in principle to those used on nearly all ships a few years ago. Though the spark is perhaps much smaller the distance is so much less,

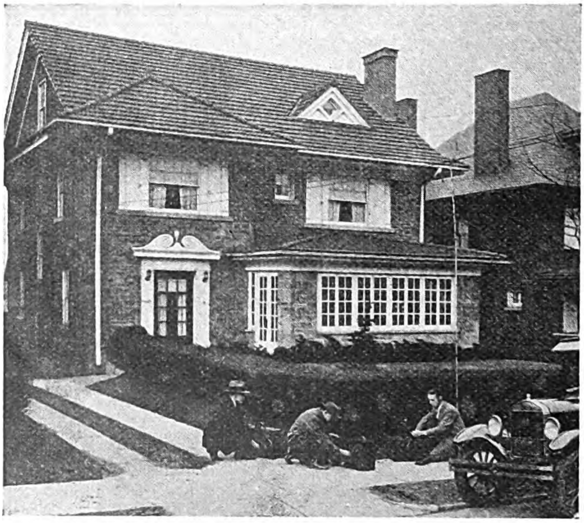

Fig. 2

and instead of signal strength diminishing as the square of the distance as in radiation over short distances, the signals are carried along the supply wires with much less diminution. This is very noticeable on some power lines where the noise level is within 40 per cent of average value at any point on the circuit.

A common idea among radio investigators was that the correct method of running down interference should be the same as that used in locating a spy's transmitter in wartime, namely to use a loop set and plot lines of maximum signal on a map. The common point of convergence should then be the source of disturbance. No such simple scheme can be used in 95 per cent of the cases. A loop picks up most of its signal by magnetic induction from the nearest lines. Whenever the

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wires turn a corner, the maximum effect turns the corner with it, thereby hopelessly confusing the investigator.

The second method tried is still in use by many investigators, and is generally known as the intensity method. In this system a careful survey is made of the suspected region and the intensity of field noticed

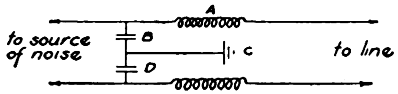

Fig. 3—Universal Filter Circuit Applicable in 80 Per Cent of All Cases.

at each point. Any apparatus found at the point of maximum intensity is then carefully inspected to determine the cause of trouble. Fig. 5 shows the dropping off in field strength as the receiver is moved away from the line carrying the disturbance.

Serious difficulties encountered with this method are:

1. Resonant points along the circuits give false clues and multiply the required work. This is shown in Fig. 6.

2. Circuits paralleling the producing circuit spread the noise over a huge area in certain instances.

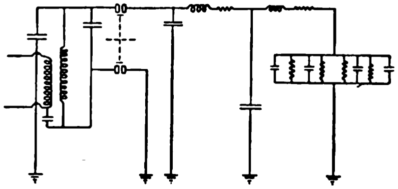

Fig. 4—Equivalent Circuit of a Precipitator.

3. The variation of the distance to ground under the wires causes corresponding variations in field intensity.

4. The presence of metal structures or masses along the line produces unaccountable variations.

5. Reflection phenomena are noted at all ground leads and at the ends of circuit branches.

To locate interference sources expeditiously you must use a combination of intensity method and direction method, and the results of experiences with similar situations.

In this method, use is made of the directional properties of the loop to differentiate between the various circuits. Thus if a house service circuit and a telephone circuit are under suspicion, a test is made at points where the wires diverge; by swinging the loop around

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the noise-producing circuit is indicated by the line of maximum intensity. This is useful in all situations regardless of the number of lines present, and fails only on the border between territory effected by one noise-producing fault and that effected by a similarly toned source.

A knowledge of the peculiarities of radio interference as learned from extensive experience in this field is most useful in making a quick determination of the trouble, and no set of rules can be formulated to take the place of such experience.

Interference travels in three ways, namely, direct radiation, conduction along metallic circuits, and combinations of conduction and radiation. Examples of the first class are ignition noises, station hetero-

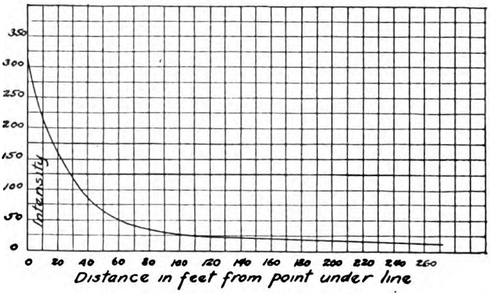

Fig. 5—Variation of Intensity Measured along Perpendicular.

dyning, and small independent plant apparatus where no incoming circuits can carry the noise out by conduction. Examples of the second class include noises carried directly to a.c. sets along the house service lines. The third class is represented by such examples as all noise originating in appliances used on the house service lines and carried along on them to be transferred by induction or radiation to the antennas of receiving sets.

Every service man should know the following simple directions for locating radio interference:

1. Remove the antenna and ground wires and notice the change in noise intensity. If the noise continues undiminished, look for set trouble. If the noise is cut down not more than 50 per cent, look for the trouble in the complainant's own house wiring and appliances as this indicates that most of the noise is entering the set by conduction. If nearly all of the noise goes when the antenna is removed, look for outside interference.

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2. Find out what electrical equipment is being used in the neighborhood. If any appliances are suspected, have them turned on and off several times while listening to the receiver.

3. If all your observations indicate that the public utility's equipment is probably the source of trouble and if you are not able to locate the exact source, then call upon the utility company for assistance.

In any city there is normally a high noise level. Extremely quiet reception from a supersensitive receiver must not be expected in such places. If sensitivity is raised each year, people must be instructed in what to expect from these supersensitive sets or general dissatisfaction results. Many dealers are losing sales by overselling the customer

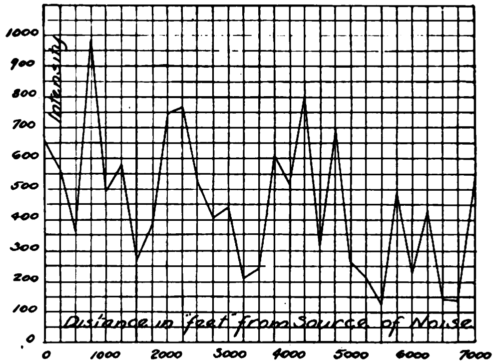

Fig. 6—Variation of Intensity Measured at Various Points along Power Circuit.

on what to expect from a new set. Sets are being turned back even after large deposits have been paid on them. Every such case is a loss to the dealer's reputation.

Some service men and salesmen are taking the easiest way out in cases where the customer is not satisfied. They say, "Call the Power Company and tell them there is a power leak, and they will be glad to know about it, as it will save lots of money by catching it before much power leaks away." The nearest thing to a power leak on overhead lines is a tree ground or broken insulator, and that amounts to a very small trickle. The results of our investigations of thousands of complaints of this nature show that in the great majority of cases the noise originates in some neighbor's electrical appliance.

A favorite expression among salesmen is "leaking transformer." This idea has caught like wildfire and extends from coast to coast. There have been transformers that leaked and gave radio interference, but they are rare enough to deserve a place in a museum.

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There would be little to gain by the work on interference if it merely proved who was to blame for the disturbances, but fortunately many of these noises can be stopped at the source or en route. Today there are a variety of filters made commercially to fit every need. Do not expect the simplest filter to take care of the more difficult forms of interference. This refers particularly to violet-ray, x-ray, electro-therapeutic machines, precipitators, and others of that class.

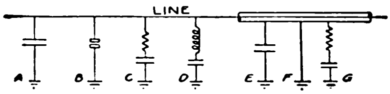

Fig. 7—Use of Impedance in Grounding.

Some idea of the complexities involved may be had by studying the equivalent circuit of a precipitator. See Fig. 4. Such equivalent circuits had to be analyzed before any progress was made in eliminating precipitator interference.

Fig. 3 shows the ordinary filter circuit which is applicable to most cases of interference.

Space will not permit a detailed explanation of the filter treatment of the many sources of interference, but a suggestion in line filter design will not be out of place at this point.

In the study of interference draining more attention should be paid to the fact that a steep wave-front surge causes reflections at the ends of circuits and also at the grounding points whether metallic grounds or through condensers.

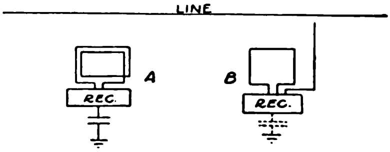

Fig. 8—Unidirectional Trouble Finder.

If the grounding impedance is equal to the high-frequency surge impedance of that portion of the line, there will be little tendency to reflection and the interference will be successfully eliminated by grounding.

Fig. 7 shows various grounds applied to a line having radio interference on it. (a) shows a direct condenser ground which produces reflection, under most conditions, which defeats the purpose of the condenser. (b) indicates grounding through small gap in low-voltage

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line. Very poor results are attained. (c and d) show respectively an impedance ground and a resistance ground, either of which may successfully eliminate the noise. However, an impedance ground may be effective for only one particular band of frequencies, while the resistance ground will be equally effective on any frequency or steepness of wave front. The remaining grounds (e, f, and g) show the same conditions applied to a cable sheath having induced interference on it.

Equipment for locating interference sources consists of superheterodynes (self-contained), low-frequency amplifiers, exploring coils, audibility meter, and miscellaneous meters. The most useful is the

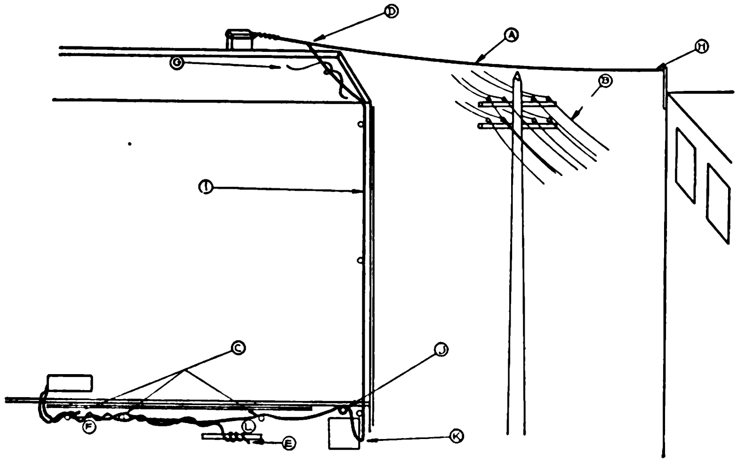

Fig. 9—Typical Faults in Antenna Installation.

suitcase model super. The set testers come in very handy on the numerous false complaints where the real trouble lies in the set, and the owners are most obstinate and troublesome. Care must be taken that this does not lead to free service work.

One piece of apparatus which will prove invaluable when once perfected is the unidirectional receiver, such as indicated in Fig. 8. The necessary development work will require much time which has not been available. Some obstacles to the use of this device at present are: bulk, lack of ground connections, and necessity of transporting delicate meters. A direct indicating audibility meter should be used in connection with this instrument.

Fig. 1 illustrates the use of a low-frequency amplifier in tracing a noise along the house service wires, and also shows a superheterodyne being used to check whether the disturbance is originating on the 110-volt supply or on the telephone lines.

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The low-frequency amplifier is very useful when near the source of trouble, as it picks up the disturbance only when at a short distance from the source and avoids the false indications met with while using a more sensitive receiver.

Fig. 2 shows various receivers being used in tracing a noise on the house service lines. Portability is essential for this work.

Fig. 9 shows antenna conditions typical of those found in half the homes today, causing not only poor signal pick-up but lack of selectivity, high interference pick-up, and the production of noises at points of contact with metal and masonry and at loose joints and clamps.

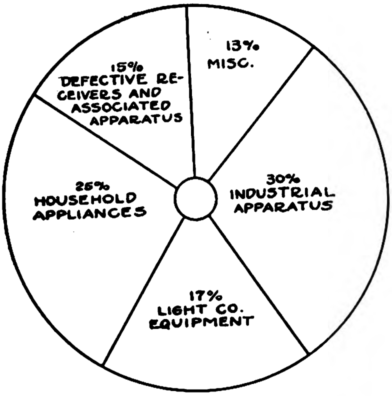

Fig. 10—Responsibility Classification.

(a) Total length of antenna and ground lead is double the correct value in the average installation.

(b) Many antennas are found passing over or under lines carrying current at high voltages. This is extremely dangerous to life and property. A broken wire or accidental contact during a storm or at the time of installation may mean death to all touching the wire or the radio.

(c) Antenna lead-ins laid over various pipes and brackets.

(d) Unsoldered and corroded joints.

(e) Ground wire loosely wrapped around corroded pipe or loose ground clamp.

(f) Lead-in and ground wire wrapped together.

(g) Lead-in tangled with loose wires or scraping against metal scraps or spouting.

(h) No insulators.

(i) Down lead along drain pipe.

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(j) Turn taken around pipe to hold wire in place.

(k) Against damp masonry.

(l) Parallel to 110-v wires and to telephone lines.

Fig. 10 shows the ultimate classification of a year's interference cases brought to a power company as their trouble. This of course does not mean that the power company was responsible for 17 per cent of all the noise in that territory, but that of all the cases where claim was made that the cause was without doubt the company's, only 17 per cent actually turned out to be such.

Radio interference work is at present being carried on mainly by the power companies to promote good will and to clear all trouble of their own before it comes to the public attention.

It is a pioneering project in several ways, and consequently is looked upon with certain misgivings by conservatives. The policy which will be followed in this work will depend largely upon the attitude of the public as given expression in letters of commendation, the public press, and in requests for this service.

This field will offer unlimited opportunities for research and constructive development when the public and the affected manufacturers realize the possibilities of improved reception conditions.

Conclusions

It is urgent that all companies or groups interested in any phase of radio should aid in the investigation and elimination of radio interference.

Many lines of investigation are open in this work, such as

(a) Unidirectional source finders.

(b) Substantial and reliable audibility meters of the indicating type, possessing high sensitivity.

(c) More thorough filter design study, with special attention to draining noises from power lines, trolley wires, and feeders and telephone lines.

(d) The manufacture of "interference proof" electrical appliances.

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Discussion on
RECEPTION EXPERIMENTS IN MT. ROYAL TUNNEL*

A. S. Eve, W. A. Steel, G. W. Olive, A. R. McEwan, and J. H. Thompson

Carl R. Englund:[1] During the delivery of his paper, Dr. Eve referred several times to radio methods of ore prospecting, and it has occurred to me that the attenuation in the ground which he reports as observed for ordinary radio signals is not necessarily the same attenuation as that which the ore prospector would encounter in applying radio to subterranean ore finding. This point is worth making even if, as the computations show, the attenuations of different wave motion types are not radically different.

In subterranean ore finding it would seem that ordinary transverse waves are the preferable type, such for example as would best be produced and observed by using buried antennas. In these waves the electric and magnetic vectors are not in time phase[2] but are at right angles to the Poynting vector, and the only serious difference from ether waves is the presence of an attenuation factor. We therefore express these waves as either the real or the imaginary part of the complex expression

$$E = E_0 e^{-\delta r + i\omega(t - r/V)} \quad \text{where}$$

$$\delta_T = \frac{2\pi\sigma}{c} \cdot \sqrt{\frac{\mu}{\epsilon}} \cdot K \qquad K = \sqrt{\frac{2}{\sqrt{1+\left(\frac{2\sigma}{f\epsilon}\right)^2}+1}}$$

$$V = \frac{c}{\sqrt{\epsilon\mu}} \cdot K$$

ϵ = dielectric const.
σ = conductivity
μ = permeability
c = vel. of light

$$f = \frac{\omega}{2\pi}$$

and for small conductivities or high frequencies the factor K is nearly unity or δ independent of the frequency.

Ordinary radio transmission is another type of wave motion, a longitudinal one in fact. The equiphase surfaces no longer coincide with the equiamplitude surfaces, and there is attenuation in directions both along and perpendicular to the ground. In a recent discussion Fry[3] has broadly treated these wave motions; they have been termed by him "hybrid" waves. That the attenuation along the direction of propagation of transverse waves should be the same as that perpendicular to the ground in a hybrid wave is neither necessary nor evident and a distinction is clearly to be made.

The simple Cohn-Zenneck[4] theory of transmission along a material boundary is based on several plausible assumptions which result in the differential equation for the magnetic field as

$$\frac{\partial^2 H_y}{\partial Z^2} + \frac{\partial^2 H_y}{\partial X^2} = \frac{i\omega\mu(4\pi\sigma + i\omega\epsilon)}{C^2} H_y,$$

the electric field being derived via Maxwell's equations from the magnetic field satisfying this equation. The plane material surface boundary is here the x, y

* Proc. I. R. E. **17**, 347; February, 1929. Presented before New York meeting of the Institute, December 5, 1928.
[1] Bell Telephone Laboratories, Cliffwood, N. J.
[2] Because of the earth's conductivity.
[3] Fry, *Jr. Opt. Soc. Amer.* **15**, 137, 1927; **16**, 1, 1928.
[4] Cohn, "Das Elektromagnetische Feld" (book), p. 449, 1900; Zenneck, *Ann. d. Phys.*, **23**, 846, 1907.

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plane, and the fields are assumed independent of the y coordinate. These are the geometrical settings, the two assumptions which have been made are: the y component of the electric field is non-existent and the time factor enters as $e^{i\omega t}$. The simplest solution of the differential equation is

$$H_y = H_{0y} e^{mx+nz}$$

with $m^2+n^2 = i\omega\mu/C^2\,(4\pi\sigma + i\omega\epsilon)$ as the only condition placed upon the unknowns m and n. When, however, we consider the boundary conditions at $z=0$, we obtain additional equations making m and n determinate. We can thus obtain an expression for the electric field in the material medium as the product of an amplitude factor and a combination amplitude and phase term, this latter being a simple exponential function of both x and z. The electric field vector is of course not vertical, and the wave motion is not a transverse one as already stated.

When $\mu = 1$ and the media are air and leaky dielectric, respectively, the exponential amplitude phase factor in the z direction is

$$e^{\pm \omega/c\sqrt{\epsilon}\sqrt{\frac{\left(i\frac{2\sigma}{f\epsilon}-1\right)^2}{\left(i\frac{2\sigma}{f\epsilon}-1-\frac{1}{\epsilon}\right)}}}$$

and the attenuation coefficient itself is the real part of the complex exponent or

$$\delta_H = \frac{\omega}{c}\sqrt{\epsilon}\sqrt{\frac{\left(\frac{4\sigma^2}{f^2\epsilon^2}+1\right)}{\sqrt{\frac{4\sigma^2}{f^2\epsilon^2}+\left(1+\frac{1}{\epsilon}\right)^2}}}\,.$$

It may be remarked that the existence of a physical counterpart of the simple Cohn-Zenneck solution is not provable by any mathematical device; its selection is moreover an act of judgment by the theorist and must be checked by physical tests. However, no transverse wave can satisfy the differential equation, and more general solutions can be built up from the simple one so that its use is doubtless legitimate.

The values of δ_T and δ_H, the attenuation coefficients for the transverse and hybrid waves, respectively, are thus different functions of the frequency and it is of interest to calculate them in particular cases. I am indebted to Mr. Sterba, of the Bell Telephone Laboratories, Inc., for the following values of earth constants, measured at broadcasting frequencies at Deal Beach, N. J.

Condition	ϵ	ohms per. cm cube	σ
Dry earth	30	600,000	1.5×10^{6}
Wet earth	80	6,000	1.5×10^{8}

and from these the table follows, the attenuations being per cm distance in every case.

	Dry Earth		Moist Earth	
Freq.	δ_T	δ_H	δ_T	δ_H
10^3	0.81×10^{-5}	1.147×10^{-5}	0.81×10^{-4}	1.15×10^{-4}
10^4	2.44×10^{-5}	3.64×10^{-5}	2.56×10^{-4}	3.62×10^{-4}
10^5	5.21×10^{-5}	13.4×10^{-5}	8.0×10^{-4}	11.47×10^{-4}
10^6	5.72×10^{-5}	113×10^{-5}	22.5×10^{-4}	36.9×10^{-4}
10^7	5.73×10^{-5}	1130×10^{-5}	34.5×10^{-4}	192.9×10^{-4}
10^8	5.73×10^{-5}	11300×10^{-5}	35.1×10^{-4}	1861×10^{-4}

Evidently as the conductivity increases the disparity between δ_T and δ_H decreases, but the frequency change of δ_H remains more pronounced than that of δ_T, and hence observations of the type reported by Professor Eve on the hybrid wave of ordinary radio transmission at and near the surface of the earth do

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not apply directly to the electromagnetic wave of transverse type which should be used for ore surveying.

A. S. Eve:[5] It is certainly true that there is marked dissimilarity between underground radio detection and ore prospecting, even when the same frequency is used.

In searching for ore by electrical methods a horizontal loop, of well insulated cable, is placed on the ground, or a vertical loop is supported on a frame. An alternating current is passed round the loop, and the frequency may be in the audible range, about 500 cycle, or of radio range, about 30 kc. The current in the loop induces an electromotive force around a good conductor underground, if not too far from the loop. This conductor reradiates energy which must be detected by a suitable detector on the surface. The fact that the loop is near and almost above the conductor, the fact that this conductor has to affect the field, and the fact of the measurement of the reradiation or secondary disturbance certainly differentiates geophysical prospecting from mere underground radio detection.

Both schemes do, however, involve the question of penetration of rocks by electromagnetic waves of varied frequencies under conditions of many different types. Mr. Englund's discussion is useful in calling attention to the difficulties surrounding this subject. In the original paper by Major Steel no reference was made to ore detection, but in my lectures at New York I certainly alluded to the subject of ore detection, which indeed first directed my attention to the absorption of radio by rocks.

[5] Director of Physics Dept., McGill University, Montreal, Canada.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 5 *May, 1929*

Discussion on

FACSIMILE PICTURE TRANSMISSION*

V. Zworykin

Alfred N. Goldsmith:[1] This evening there has been presented an interesting paper on "Facsimile Picture Transmission." Those of us who can remember back to the early days of radio communications are aware that spark transmission was the order of that already remote day. Modulation, as such, was a strange term to radio engineers, because modulation implies taking something which was rather steady and then impressing something else upon it. But there was not anything so definitely dependable and constant about a spark gap; it broke down in spasmodic and erratic fashion with the utmost violence and considerable noise, and gave rise to trains of damped waves which by the most charitable interpretation could not be regarded as a modulation of a continuous wave. The effect of such waves on the receiving set and the patience of the person trying to avoid interference was very marked. Thus modulation in those days was practically without significance as a radio engineering term. Then came the quenched spark transmitter and sometimes wave trains from such transmitters even overlapped, although with random phase relationship between the successive groups of waves and therefore still without any very clear meaning to the term "modulation" as employed in present day practice.

Finally, there came the era of the continuous-wave transmitting set, of the telegraph type first, where modulation or control of a continuous wave was accomplished slowly but surely (very slowly over the average circuit, and sometimes none too surely).

As the years went on, radio telephony, with a more complex form of modulation, came into existence and the problems of control of the transmitter and adequate reception of the extended sidebands corresponding to this form of modulation required a period of years for adequate development.

Today we stand on the threshold of the newest, most remarkable and unusual types of modulation, the modulation of electromagnetic waves to correspond to the still picture or to the changing contours of a motion picture. We are endeavoring to impress upon the flying wave the likeness of persons or things, either stationary or mobile. These new fields are engaging the serious attention of radio workers because clearly facsimile transmission has something of the quality of intimate contact with the original which telephony has not and which is lacked, necessarily, by the more cold telegraphically-transmitted word. Television requires even more complicated modulation, and we are beginning to think in terms of sidebands which would have been truly disconcerting in their width to the telegraphic communication engineer of a few months ago. Nevertheless much has been accomplished and a description of a substantial and practical contribution to the subject is presented by Dr. Zworykin.

M. B. McCullough:[2] What applications does a system of this type have?

V. Zworykin:[3] I would say that the application of a system of this type would be to any sort of facsimile transmission; in other words, it may be used to transmit and broadcast newspaper pictures, transmit commercial telegraphs, transmit orders from one factory branch to another, and so forth.

N. E. Bonn: How much power must the tuning fork supply the motor to keep it in synchronism with the fork?

V. Zworykin: As much power as possible, and the more the better. Ten watts is sufficient to keep the picture quite stable, but this is about the lower limit.

N. E. Bonn: You have constant power and constant voltage?

V. Zworykin: Not necessarily. We are using the regular power supply.

* Proc. I. R. E., **17,** 536; March, 1929. Presented before New York meeting of the Institute, January 2, 1929.

[1] Chief Broadcast Engineer, Radio Corporation of America, New York, N. Y.

[2] Sales Department Radio Corporation of America, New York, N. Y.

[3] Research Department, Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pa.

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A. Nyman:[4] I notice that the method used for introducing the alternating-current component in the picture is by interrupting the light. I wonder if the alternative method of using alternating current supplied to the photo-electric cell has been considered; and if so, what are the objections to it?

V. Zworykin: Yes, it has been considered, and we have been working for a number of years with this method. It is a very alluring method, I would say, because the elimination of the light interrupter is very desirable. The trouble is that you have to use quite a high frequency, at least 3000 cycles. In this case the capacity of the photo-electric cell becomes quite important, for at such frequencies the capacity leakage is relatively large.

E. C. Ballentine:[5] I would like to ask Dr. Zworykin if he has actually found that the helium lamp is as good as the neon? Since we developed our neon lamps about a year ago, we have been experimenting with different types, and it has been our experience that helium is actually inferior to neon, for the reason that neon has some violet bands in the spectrum which are hardly visible, but very actinic.

V. Zworykin: I admit that this question is still not quite settled. I think the spectrum of the glow discharge in this kind of tube depends not only on the gas used but also on the character of the glow discharge. You can make the constricted glow discharge by using either a positive column discharge or a negative column discharge. In two different cases there will be two different results, and since no close investigation was made, I cannot give you a definite answer on that.

E. C. Ballentine: I would like to make just one remark, if I may. I would like to impress on the Institute at this time a feature of facsimile which Dr. Zworykin has not mentioned in his paper and which I think is quite important. That is one thing about facsimile that the General Electric Company has been experimenting with for the last two years—the radio circuit. We have found that the short waves are very variable things to work with. We find that no matter how good our terminal apparatus may be, or how perfect it may be, the major problem still remains in radio of getting the high speed telegraphic signals across and through the ether. I think that still remains the major problem in radio facsimile work.

Alfred N. Goldsmith: There can be no question, of course, that once the terminal apparatus has been brought to a degree of perfection greater than that imposed by variations of attenuation throughout the transmission band, any improvement of terminal apparatus is merely something to be stored up against the happier day in the future when the radio circuit is smoothed out or improved in its transmission characteristic.

Taking a transoceanic circuit capable of carrying, say 100 cycles or 200 cycles per second, it obviously would not be of any direct commercial advantage today to be able to receive 1000- or 10,000-cycle modulation in a receiver or to send it out from a transmitter, because such modulated waves would be too badly mangled in transmission. However, the inherent advantages of the improved equipment will appear as soon as the radio circuit is correspondingly improved. Fortunately we are gradually improving our radio links as well. The limitation of some radio circuits is a feature which is neglected at times, and which frequently accounts for the amazing results obtained both in so-called facsimile and so-called television transmissions where the transmitting medium consists of an extremely accommodating and very solid length wire from one room to another.

H. H. Haglund: I understand that the synchronism maintained by having the two tuning forks vibrating at a different speed is about 1 beat in 20 seconds; that is, the second receiving tuning fork is corrected once every revolution of the drum. I would like to know how this correction is applied to the fork.

V. Zworykin: The tuning fork is driven by a vacuum tube and a resistance is included in the grid circuit of this tube. Every time that the correcting impulse comes to the terminals of the resistance it changes slightly the oscillation of the tube and this corrects the tuning fork. We have found this is sufficient to correct the second tuning fork, if the frequency difference between tuning forks does not exceed about 1 beat in 20 seconds.

[4] Radio Patents Corporation, New York, N. Y.
[5] Radio Engineering Department, General Electric Company, Schenectady, N. Y.

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T. P. Dewhirst: I would like to ask what is the average life of the helium glow lamp as compared with the neon lamp, and the limitations of flash and voltage?

V. Zworykin: I do not see why there should be any difference in the life of the helium lamp as compared with the neon lamp. The limitation is mostly from the sputtering of the electrodes, which is about the same, in both cases.

T. P. Dewhirst: What power is available across the polarizing resistance?

V. Zworykin: One lumen gives about 20 microamperes through 0.5 megohm, with caesium.

T. P. Dewhirst: How about the potassium cells?

V. Zworykin: In our measurements we found the caesium cell was about 100 times as sensitive as the potassium cell.

J. E. Shannon: How does the photographic image at the receiving end compare with the original? Take, for instance, a perfect photograph at the transmitting end: how does that compare at the receiving end?

V. Zworykin: It depends, of course, on how many picture elements you transmit. As you know, for newspaper work they use about 50 lines per inch. In our case we use 64 lines per inch, and we have found it quite satisfactory. On the other hand, in some other facsimile machines they use a much higher number of lines. For instance, the General Electric engineers use 100 lines per inch and for the fine details, of course, it is better. In Europe I saw systems using 130 lines per inch. I do not see much difference from 100 lines up and believe that further increase is a waste of time and of the width of the band of transmission.

R. H. Ranger: I would just like to emphasize what Mr. Ballentine has brought out. Not only are you limited in facsimile transmission by the radio circuit, but it does seem to me that you have to make your apparatus conform to the radio circuit. In other words, there are many ways of solving any particular problem, and the one that you should select is the one that will get through the complete circuit from one point to the other with the least power through the most interference, and that is why there is such a discrepancy between some short link laboratory results and those that go through the air.

The great bugbear that we have found in all our work is the change in the level of the received signal. Most of the very high-speed facsimile test work, on the contrary, has been on a basis where the change level is very important in order to get the tones. All the work that we have done in the Radio Corporation has been on the "on and off" basis with the interpretation of the halftone values in terms of dots spaced at different distances apart, because we have found that that is the only reliable method of getting current values that will register correctly over great distances. That all means, of course, that you have to slow down the rate at which you scan, because you have to interpret your intensity values in terms of separation and length of the dot elements, and therefore your scanning speed goes down proportionately.

In regard to the number of scanning lines which you use, there again you are tied to the radio circuit. If you use 60 lines to the inch, you will get along very nicely, if you have a perfect radio circuit. On the other hand, if interference, fading, or any other irregularities enter, you will find your chances for getting through with a double number of lines per inch vastly improved. For example, on some of our work where we use 120 lines to the inch we can miss a whole line through fading or other causes and still get a perfectly legible typewritten copy, and that of course is our goal.

In regard to the pick-up system with the light and parabolic reflector, I am quite in accord that that is a novel and efficient way of picking up light from a spot. However, strange as it may seem, it may not be the most efficient way to pick up black from a spot. In other words, what we are concerned with in picture transmission is getting the greatest differentiation between black and white. Therefore, you have to pick the optical system which will give you that greatest differentiation. To get the greatest differentiation, you want as fine a pencil of light shot at the paper as possible, and you likewise want as fine a pencil of light focussed back from the paper. In other words, if you send a pencil of light straight at the piece of paper, you get diffuse reflection on all sides, and if you pick it up from all sides you will find that as your image crosses in front of that pencil of light, first the area will be darkened on

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one side and then on the other. Therefore, you may not get clean-cut differentiation as the pencil of light passes through the type or other picture matter. However, all these matters are compromises and therefore what you have to select is the one that best fits your whole proposition. The real answer of the effectiveness of this particular compromise is the excellent pictures that Dr. Zworykin has shown.

In regard to the synchronizing, again it is the proposition of realizing that you have to work imperfectly through space. If you depend upon the radio circuit to see to it that your rotating drums are in absolute step, you are just requiring that much more from your poor link through space. If you can get synchronism in any other way than by radio, therefore, it is very wise to concentrate on that other way and leave the radio channel to get the lights and shades across. At the present time, for example, we find it quite possible to maintain separate tuning forks on opposite sides of the continent, within two parts in a million and the drift which that means in any picture is certainly very slight; in fact, you would not be able to notice it. I feel that we have to concentrate on the things that fit the conditions as we meet them in the air, and while I feel that the Chairman's statement is correct, that the greater abilities of the terminal equipment will come into their own when the radio circuit is improved, nevertheless we have to take things as they are and make the most of them by adapting our terminal equipment to those conditions.

Alfred N. Goldsmith: There can be no doubt that 100 per cent modulation, and not too rapid modulation, is desirable when you are transmitting through an "ether" that has variable transmission characteristics, different zero levels for all frequencies, and a fairly high interference level.

I had occasion recently to point out that in the early days of happy simplicity in radio transmission, we believed that we were transmitting through a homogeneous, non-absorbing medium, but that we have since learned that we actually transmit through a swirling fog of varying colors, in which are imbedded numerous rapidly moving, irregularly surfaced mirrors, and crude lenses or prisms.

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BOOK REVIEW

Handbuch der Experimentalphysik. Edited by Wein-Harms. Vol. xiii, Part 2. Glühelektroden und Technische Elektronenröhren, by W. Schottky, H. Rothe, and H. Simon; p. 492, 179 figs. Akademische Verlagsgesellschaft M.B.H. Leipzig; price bound 46 RM.

This volume is in reality a collection of three monographs, each dealing with a different phase of the radio vacuum-tube problem.

The first part, entitled "Physik der Glühelektroden," and occupying well over half the volume, is written by Schottky and Rothe. The authors develop the present status of theoretical and experimental thermionics, principally in the language of the Gibbs thermodynamic potential. Following preliminary orientation chapters covering a treatment of the thermodynamic principles on which this part of the volume is based, chapters on electron vapor pressure and emission, thermionics, thermal effects of ionic emission and the surface activation problem, are among those included. A final chapter deals with phenomena not approachable by thermodynamic methods, in particular the Schottky effects. Experimental work and theory are well coordinated throughout the monograph.

The second part of the volume, a short section of less than sixty pages, presents "The preparation of the Hot Cathode," by Simon. Experimental methods are described for filament preparation and mounting for consistent behavior as an electron emitter, rather than the actual commercial practice of any particular tube manufacturer. The behavior of the different types of hot cathode and the methods of evacuation and of temperature measurement are also dealt with in this survey of the literature on hot cathode technique.

The radio engineer will wish that the third monograph were longer and more complete: "Technical Electron Tubes and their Application," by Rothe. While the two previous parts are of primary interest to physicists and to the specialist in filament development, there is found in this latter part, comprising somewhat less than a third of the volume, a review of the literature on fields more familiar to the radio engineer. Following a treatment of vacuum-tube characteristics and of circuits for their measurement come sections on amplifier, oscillator, modulator, and detector, with emphasis on circuits which have actually come into practical use. Neutralizing and feed-back arrangements in amplifiers occupy several pages. Two pages are found for the four-element tube and its circuits, and fifteen pages for the quartz

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resonator and its control of oscillating circuits. Here, as is apparently also true with other subjects covered in the monograph, the author has given a good introduction to the piezo-electric resonator, but has not had space available for the treatment of details. Developments which have not so far come to have practical importance, such as the dynatron and the magnetron, are omitted entirely.

The entire volume is well prepared. The reviewer finds a number of inconsistencies in the manner of citing periodicals, some journals being referred to in as many as three different ways. It would make for convenience in such a reference work if each page bore either the title or the number of the monograph to which it belongs.

K. S. Van Dyke

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MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The articles listed below are not obtainable from the Government. The various periodicals can be secured from their publishers and can be consulted at large public libraries.

R100. Radio Principles

R113 Barfield, R. H. The attenuation of wireless waves over towns. *Jnl.* Institution of Elec. Engrs. (London), **67**, pp.253–270; February, 1929.

(Shows that the rate of change of attenuation with wavelength of the radiated field over towns is rather great. A theoretical discussion is given by means of Sommerfeld's numerical distance and the same applied to the absorbing effect of vertically grounded conductors which exist in the towns in many forms. Also a theoretical discussion on selective absorption due to tuned antennas and their reradiation. Sommerfeld's theory seems to confirm the high rate of attenuation over towns.)

R113 Sacklowski, A. Die Ausbreitung der Elektromagnetischen Wellen. (The propagation of electromagnetic waves). Book. Published by Weidmannsche Buchhandlung, Berlin, 1929. Noted in *Experimental Wireless and Wireless Engr.* (London), **6**, p. 83; Feb., 1929.

(Review of work which was published by the author in the *Elektrische Nachrichten Technik*, 1927.)

R113 Eve, A. S., Steel, W. A., Olive, G. W., McEwan, A. R., Thompson• J. H. Reception experiments in Mount Royal Tunnel. Proc. I. R. E., **17**, pp. 347–376; Feb., 1929.

(Experiments presented which were carried out in the Mt. Royal Tunnel of Canadian National Railways at Montreal in order to determine how radio waves reach the receiving set.)

R114 Yokoyama, E. and Nakai, T. A note on the directional observations on grinders in Japan. Proc. I. R. E., **17**, pp. 377–79; Feb., 1929.

(Direction of grinders taken in Japan during 1927 showed diurnal and seasonal changes in the direction of atmospherics.)

R125.6 Transmitting aerials (editorial). *Experimental Wireless and Wireless Engr.* (London), **6**, pp. 59–61; Feb., 1929.

(Discusses the features of the beam system developed by A. Meissner ($f = 15000$ kc) and that due to P. P. and T. L. Eckersley and H. L. Kirke for the broadcast range. Shows the importance of emphasizing the ground ray.)

R125.6 Eckersley, P. P. and T. L., and Kirke, H. L. The design of transmitting aerials for broadcasting stations. *Experimental Wireless and Wireless Engr.* (London), **6**, pp. 86–92; Feb., 1929.

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(It is brought out that the radiation along the tangent of the earth's surface should be emphasized and any appreciable intensities upwards suppressed. It is shown that this can be accomplished with high antennas working with more than quarter wavelength distribution (for instance full wavelength distribution utilizing phasing coils). The attenuation due to ground absorption is also studied by means of Sommerfeld's formula.)

R132 Garton, C. G. and Lucas, G. S. An apparatus for the projection of frequency output characteristics. *Experimental Wireless and Wireless Engr.* (London), **6**, pp. 62–70; Feb., 1929.

(Description of a system for studying frequency amplitude characteristics of audio-frequency amplifiers.)

R132 Nelson, J. R. Circuit analysis applied to the screen grid tube. Proc. I. R. E., **17**, pp. 320–338; Feb., 1929.

(General radio-frequency circuit theory is discussed in this paper. Amplification equations for impedance and transformer coupling using an untuned primary whose period is above the highest frequency considered are derived and discussed for the case of a screen-grid tube.)

R133 Hollmann, H. E. On the mechanism of electron oscillations in a triode. Proc. I. R. E., **17**, pp. 229–251; Feb., 1929.

(Considers the different types of electron oscillations which occur in the Barkhausen and Kurz retarding field of a triode having a high positive potential applied to its grid, and zero or small negative potential applied to its plate.)

R134.4 van der Pol, B. The effect of regeneration on the received signal strength. Proc. I. R. E., **17**, pp. 339–346; Feb., 1929.

(Theory of the effect of regeneration using the solution of a non-linear differential equation. Experimental verification of the theory given.)

R144 Cockcroft, J. D. Skin effect in rectangular conductors at high frequencies. *Proc.* Royal Soc. A (London), **122**, pp. 533–542; Feb. 4, 1929.

(Deduction of formulas for the high-frequency resistance of conductors with rectangular cross section.)

R200. Radio Measurements and Standardization

R210 Hull, L. M. and Clapp, J. K. A convenient method for referring secondary frequency standards to a standard time interval. Proc. I. R. E., **17**, pp. 252–271; Feb., 1929.

(Method is described for obtaining a convenient low frequency from a high-frequency standard by means of harmonic control of distorted wave oscillators (multivibrators).)

R210 Hall, E. L. A system for frequency measurements based on a single frequency. Proc. I. R. E., **17**, pp. 272–282; Feb., 1929.

(Method is described which is suitable for calibration of either piezo oscillators or frequency meters in terms of an accurately standardized temperature controlled piezo oscillator.)

R300. Radio Apparatus and Equipment

R320 Landon, V. D. A multiple receiver antenna system. *Radio Broadcast,* **14**, pp. 291–293; March, 1929.

(System for apartment houses.)

R333 Weaver, K. S. A high-power output tube—The 250. *Radio Broadcast,* **14**, pp. 329–330; March, 1929.

(Characteristics of tube.)

R333 What is a good tube? *Radio Broadcast,* **14**, pp. 335–45; March, 1929.

(Data given on the Cunningham, Sylvania, Ceco and Raytheon tubes.)

R343 Hull, R. A. Improving short wave phone reception. *QST,* **13**, pp. 9–20; March, 1929.

(Description of a modern superheterodyne for short-wave phone, code, and general broadcasting.)

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R376.3 Bostwick, L. G. Acoustic considerations involved in steady state loud speaker measurements. *Bell System Tech. Jour.*, **8**, pp. 135–158; January, 1929.

(Describes certain difficulties encountered in acoustic measurements of performance of loud speakers. Data given for two representative types of loud speakers when measured in space free from reflections and when measured under varying conditions in a special acoustic laboratory.)

R381 Siegmund, H. O. The aluminum electrolytic condenser. *Bell System Tech. Jour.*, **8**, pp. 41–63; January, 1929.

(Action of the aluminum electrolytic condenser and method of avoiding the rectification effect. Shows that the life of such condensers is high, some having been in use for seven years.)

R500. Applications of Radio

R521 Pickerill, E. N. A modern radio aircraft installation. *Radio Engineering*, **9**, pp. 46–48; Feb., 1929.

(Description of new transmitter and receiver designed for airplane communication.)

R521 Drake, F. H. An aircraft radio receiver for use with rigid antenna Proc. I. R. E., **17**, pp. 306–319; February, 1929.

(Outline is given of physical and electrical requirements of an aircraft receiver suitable for reception of radio beacons and telephone service on a small rigid antenna. Describes the design of a special unicontrol receiver calculated to fulfill these requirements.)

R521 Pratt, H. and Diamond, H. Receiving sets for aircraft beacon and telephony. Proc. I. R. E., **17**, pp. 283–305; February, 1929.

(Design details for three receiving sets of slightly different types with characteristic and performance curves are discussed. Reprint of Bureau of Standards Research Paper No. 19, Bureau of Standards Journal of Research, October, 1928.)

R526.4 Alexanderson, E. F. W. Height of airplane above ground by radio echo. *Radio Engineering*, **9**, pp. 34–35; Feb., 1929.

(Preliminary work on a radio altimeter.)

R531.2 Schedule of short wave programs. Radio Broadcast, **14**, p. 298; March, 1929.

(List of short-wave stations.)

R800. Non-Radio Subjects

534 MacKenzie, D. Sound recording with the light valve. *Bell System Tech. Jour.*, **8**, pp. 173–183; Jan., 1929.

(The light valve of Wente is described. It is an electromagnetic shutter made up of a loop of duralumin tape formed into a slit at right angles to a magnetic field. Amplified currents from a microphone flow in the duralumin and open and close the loop. An incandescent filament is focussed on the light valve the slit of which is focussed by a lens on the sound negative film. When sound currents flow the film receives a varying exposure.)

534 Norris, R. F. The acoustimeter. *Radio Engineering*, 9, pp. 36–37; February, 1929.

(An electrical means for measuring sound intensity.)

621.319.2 Guillemin, E. A. and Glendinning, W. On the behavior of networks with normalized meshes. Proc. I. R. E., **17**, pp. 380–393; Feb., 1929.

(Theory of normalizing meshes in electrical networks is verified and illustrated by examples and figures relating to two and three-mesh circuits.)

621.385 Osborne, H. S. The principles of electric circuits applied to communication. *Bell System Tech. Jour.*, **8**, pp. 3–20; Jan., 1929.

(Discussion of method of presenting fundamental electrical principles. Outline of problems arising in application of electric principles to telephone systems.)

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Wireless Telegraph and Telephone Company, Berlin-London (Poulsen system) 1907–09; chief engineer, Telefunken Company, 1910; acting member of the management of the Transradio A. G. für drahtlosen Überseeverkehr, 1921–.

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The Institute of Radio Engineers

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Entered as second class matter at the Post Office at Menasha, Wisconsin.

Acceptance for mailing at special rate of postage provided for in the Act of February 28, 1925, embodied in paragraph 4, Section 412, P. L. and R. Authorized October 26, 1927.

Published monthly by

THE INSTITUTE OF RADIO ENGINEERS, INC.

Publication office, 450–454 Ahnaip Street, Menasha, Wis.

BUSINESS, EDITORIAL, AND ADVERTISING OFFICES,

33 West 39th St., New York, N. Y.

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SUGGESTIONS FOR CONTRIBUTORS TO THE PROCEEDINGS

Preparation of Paper

Form—Manuscripts may be submitted by member and non-member contributors from any country. To be acceptable for publication manuscripts should be in English, in final form for publication, and accompanied by a summary of from 100 to 300 words. Papers should be typed double space with consecutive numbering of pages. Footnote references should be consecutively numbered and should appear at the foot of their respective pages. Each reference should contain author's name, title of article, name of journal, volume, page, month, and year. Generally, the sequence of presentation should be as follows: statement of problem; review of the subject in which the scope, object, and conclusions of previous investigations in the same field are covered; main body describing the apparatus, experiments, theoretical work, and results used in reaching the conclusions and their relation to present theory and practice; bibliography. The above pertains to the usual type of paper. To whatever type a contribution may belong, a close conformity to the spirit of these suggestions is recommended.

Illustrations—Use only jet black ink on white paper or tracing cloth. Cross-section paper used for graphs should not have more than four lines per inch. If finer ruled paper is used, the major division lines should be drawn in with black ink, omitting the finer divisions. In the latter case, only blue-lined paper can be accepted. Photographs must be very distinct, and must be printed on glossy white paper. Blueprinted illustrations of any kind cannot be used. All lettering should be $^3/_{16}$ in. high for an 8 x 10 in. figure. Legends for figures should be tabulated on a separate sheet, not lettered on the illustrations.

Mathematics—Fractions should be indicated by a slanting line. Use standard symbols. Decimals not preceded by whole numbers should be preceded by zero, as 0.016. Equations may be written in ink with subscript numbers, radicals, etc., in the desired proportion.

Abbreviations—Write a.c. and d.c., kc, μf, $\mu\mu$f, emf, mh, μh, henries, abscissas, antennas. Refer to figures as Fig. 1, Figs. 3 and 4, and to equations as (5). Number equations on the right in parentheses.

Summary—The summary should contain a statement of major conclusions reached, since summaries in many cases constitute the only source of information used in compiling scientific reference indexes. Abstracts printed in other journals, especially foreign, in most cases consist of summaries from published papers. The summary should explain as adequately as possible the major conclusions to a non-specialist in the subject. The summary should contain from 100 to 300 words, depending on the length of the paper.

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Proofs—Galley proof is sent to the author. Only necessary corrections in typography should be made. *No new material is to be added.* Corrected proofs should be returned *promptly* to the Institute of Radio Engineers, 33 West 39th Street, New York City.

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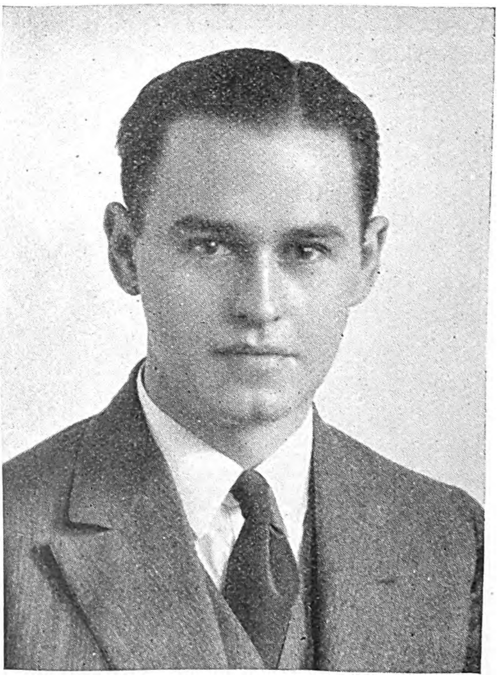

LEWIS M. HULL
Member of the Board of Direction, 1929

Lewis M. Hull was born on February 27th, 1899 at Great Bend, Kansas. He graduated from the University of Kansas in 1917 and immediately entered the Radio Section of the Bureau of Standards. He was associated with the Bureau as assistant physicist, and later as associate and consulting physicist, at intervals until 1922. In 1922 he received the degree of Ph.D. in physics from Harvard University.

In 1922 Dr. Hull joined the Radio Frequency Laboratories of Boonton, N. J. as director of research. He is now vice president of that company.

Dr. Hull has served on a number of committees of the Institute, and was appointed Chairman of the Committee on Broadcasting in 1928. He was appointed to membership on the Board of Direction for 1929.

He became a Junior member of the Institute in 1917; Associate in 1919; Member in 1927; and Fellow in 1928. He is a Member of the American Institute of Electrical Engineers and the American Physical Society.

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INSTITUTE NEWS AND RADIO NOTES

May Meeting of the Board of Direction

The May meeting of the Board of Direction was held on the first of the month in the office of the Institute in New York City. The following were present: A. Hoyt Taylor, John M. Clayton, Arthur Batcheller, Alfred N. Goldsmith, C. M. Jansky, Jr., R. H. Marriott, and L. E. Whittemore.

The following were transferred or elected to higher grades of membership in the Institute: elected to the Fellow grade: John A. Slee; transferred to the Member grade: James D. Durkee, Edwin B. Dallin, S. S. MacKeown, E. E. Cordrey, Edwin L. Powell, Paul D. Andrews, Leslie F. Curtis, E. G. Shalkhauser, F. L. Schoenwolf, O. W. Pike, John B. Brady; elected to the Member grade: Loy E. Barton, Lloyd A. Briggs, Mortimer Frankel, T. Kono, H. W. Dreyer, Frank N. Calver, E. J. Emery, James K. Nicholson, Alan G. Dryland, J. G. A. Sutherland.

One hundred and sixty-one Associate members and ten Junior members were elected.

The Board appointed C. W. Latimer as alternate to Captain Guy Hill as delegate to the World Engineering Congress to be held in Tokio in October. Captain Hill and Mr. Latimer will present the symposium paper prepared by six members of the Institute.

New York Meetings Suspended During July and August

Carrying out the usual custom there will be no New York meetings during the months of July and August. The Institute office will, however, be open for the transaction of business.

Most of the sections also will suspend their meeting activities during these two months.

Institute Meetings

Atlanta Section

Dr. W. G. Cady presented a paper, "Piezo-electric Crystals as Frequency Standards," before the Atlanta section on April 22nd. The meeting was held in the Physics Laboratory, Georgia Institute of Technology, Atlanta, Georgia, presided over by P. C. Bangs, acting chairman. There were in attendance thirty-four members and guests.

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This paper dealt with the crystal itself, classified as to different types and crystal structure, and stressed the point that the symmetry and non-symmetry of the crystal structure was important.

The crystalline structure of Rochelle Salts was explained for the purpose of showing why Rochelle Salts crystals were not used as frequency standards. While such crystals oscillate freely, the structure itself is not adaptable to any great amount of pressure or voltage.

A detailed description was given in regard to the vibration of quartz plates. The important point brought out here was that a vibrating quartz plate takes the shape of a parallelogram when oscillating. This change of shape takes place with each cycle.

Different methods of cutting crystals were explained in detail. Differentiation was made between the Curie type crystal which has a wavelength of about 105 m per mm and the later type thirty-degree crystal which has a wavelength of about 150 m per mm. The later type oscillates more rapidly.

A general discussion followed the presentation of the paper.

Cleveland Section

Dr. M. Luckiesh, director of lighting research, National Lamp Works of General Electric Company, was the speaker at the Cleveland section meeting, which was held April 26th in the Physics Building, Case School of Applied Science, Cleveland, Ohio, and presided over by Bruce W. David, chairman of the section. There were forty-two members and guests in attendance. The paper presented by Dr. Luckiesh was entitled, "Radio Messages of Men, Atoms, and Stars." The speaker stressed the similarity of man's radio messages and the messages sent out in the form of radiated energy by everything in the universe from atoms to stars. The receiving set for these radiated messages is the spectroscope and the message is received through the eye and not the ear. Man's vision limits the number of wavelengths he receives to a relatively narrow band. Perhaps some day man will be able to see life on other planets and even see images from the earth reflected by some other planet but delayed by hours, days, or years, depending upon the distance the reflected light has travelled. A general discussion followed the presentation of this paper.

The Cleveland section held a meeting at the Ohio Insulator Co., Barberton, Ohio, on May 3, 1929. D. Schregardus, chairman, presided. Fifty-three members attended the meeting.

A. O. Austin, chief engineer of the Ohio Insulator Co., presented a paper, "High Voltage." A high voltage demonstration was made

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whereby high voltage testing methods were demonstrated in a million-volt indoor laboratory and a number of fundamental principles of high voltage discharges were discussed. In the outdoor laboratory demonstrations were made of actual power line construction using voltages as high as two or three million, as artificial lightning. The demonstrations were extremely impressive and represented the highest development in designing and testing high voltage equipment. A general discussion followed the presentation of the paper.

Detroit Section

F. A. Firestone, of the Physics Department of the University of Michigan, was the speaker at the January 19th meeting of the Detroit section. E. D. Glatzel, chairman, presided, and forty members and guests attended the meeting.

The paper, "Technique of Sound Measurements," by F. A. Firestone, explained the classical experiments of Prof. D. C. Miller. The recording of the wave form of sounds was explained with reference to the methods used for analysing the records into harmonic components and of synthesizing the components into the original wave form. The ability of the ear to distinguish intensities of tones of the same and different frequencies was discussed. Apparatus used in connection with the detection of undesirable noises in automobiles in motion was shown and demonstrated. This apparatus indicated the intensity of the total noise. It can be adjusted so as to indicate the intensity of a sound of a certain pitch separately from other pitches which may be present.

The annual election of officers was held with the following results: A. B. Buchanan, chairman; Lewis N. Holland, vice-chairman; Walter R. Hoffman, secretary-treasurer.

The Detroit section held a meeting February 15th in the auditorium of the Michigan Bell Telephone Building, Detroit, Michigan, presided over by A. B. Buchanan, chairman. Two hundred and seventy-five members and guests attended.

C. L. Carrington, division superintendent of Installation of Electrical Research Products, Inc., presented a paper, "Sound and Talking Picture Problems and Equipment." The two systems of recording and reproducing the sound of talking motion pictures developed by the company employing him were described, as well as some of the means of eliminating or minimizing bad acoustical conditions encountered in some of the theatres in which the equipment is being installed. A motion picture and a number of the units of equipment comprising the reproducing apparatus were exhibited.

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Proposed Lehigh Valley Section

The third organization meeting of the proposed Lehigh Valley section was held April 12th in the Science Bldg., Muhlenberg College, Allentown, Pa. Carlton F. Maylott, temporary chairman, presided. Thirty one members and guests attended the meeting.

Carlton F. Maylott presented a paper, "The Electrical Reproduction of Speech and Music." The history of the talking machine and other recorded sound reproducing mediums and the various steps in their development were treated. A detailed description of the several commercial systems of sound motion pictures was presented and illustrated with blackboard diagrams. By means of an electrically driven turntable and a power amplifier a group of special disk recordings of the Bell Telephone Laboratories were reproduced through an auditorium type of moving-coil speaker, demonstrating the effect on speech and music of successively suppressing various frequencies in both the upper and lower tonal ranges. The committee was fortunate in obtaining for this meeting practically every distinctive type of loudspeaker marketed and this afforded a very interesting demonstration of the evolution of the radio reproducer, from the original crude horn type to the latest and most modern development of the reproducer used in conjunction with sound motion picture presentation.

Los Angeles Section

On April 15th the Los Angeles section held their meeting in the Elite Cafe, Los Angeles, Cal., presided over by T. C. Bowles, vice chairman. There were forty members and guests in attendance.

E. H. Butler presented a paper on "Vacuum Tubes and Their Applications to Radio." The paper covered especially the use of the new 45-type power tube as well as the use of the four-electrode tube as an r-f amplifier and power detector feeding directly into the power tube. The various merits of power tubes and also detectors of various types were discussed. There was quite a discussion regarding the merits of more r-f amplification with only one audio stage or the more common type of two-stage audio amplifier with conventional detector and r-f amplifier.

Philadelphia Section

The Philadelphia section held a meeting on April 25th at the Franklin Institute, Philadelphia, Pa. J. C. Van Horn, chairman, presided. One hundred and sixty members attended the meeting.

C. Francis Jenkins presented a paper, "Latest Phases of Television."

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Messrs. Synder, Miller, Arnold, Van Horn, and others participated in the discussion which followed.

PITTSBURGH SECTION

The Pittsburgh section held their regular meeting in the Fort Pitt Hotel, Pittsburgh, Pa. on March 19th. L. A. Terven, vice-chairman of the section, presided. The paper, "Recent Developments in Super Heterodyne Receivers," was presented by G. L. Beers. This paper was published in the March issue of the PROCEEDINGS.

Messrs. Allen, Wallace, Terven, Mag, Bricker, Sunnergren, and McKinley participated in the discussion which followed.

The Pittsburgh section held a meeting April 16th in the Fort Pitt Hotel, Pittsburgh, Pa. W. K. Thomas, chairman, presided. There were present twenty-nine members and guests.

Dr. G. A. Scott, of the University of Pittsburgh, presented a paper, "Mechanical Analogies of Electrical Conditions." This paper deals entirely with electro-mechanics. A comparison of the fundamental units of mechanical engineering was made to the fundamental units of electrical engineering through a demonstration of their direct similarity. A derivation of fundamental circuit formulas of electrical engineering was made by a substitution method of well known mechanical formulas.

Messrs. Terven, Hallie, Boardman, McKinley, Smith, Bordini, and Sunnergren participated in the discussion which followed.

The election of officers was held with the following results: L. A. Terven, chairman; A. J. Buzzard, vice-chairman; T. D. Cunningham, secretary; A. Mag, treasurer.

SAN FRANCISCO SECTION

The San Francisco section held a joint meeting with the local section of the American Institute of Electrical Engineers, April 10th, at the Engineers' Club, San Francisco, Calif. D. B. Dexter, chairman of the San Francisco section of the American Institute of Electrical Engineers, presided. There were one hundred and sixty members and guests in attendance.

Harry R. Lubcke presented a paper, "Design Equations for Vacuum-Tube Voltmeters." This very interesting paper gave formulas for calculating rapidly the plate external resistor value, the *C* bias value and plate voltage value for a vacuum-tube voltmeter of any given scale range.

Elmer H. Fisher presented a paper, "Voltage Surges in Audio-Frequency Apparatus." It was shown by means of an oscillograph

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and an inverted vacuum-tube voltmeter that voltages up to seven hundred volts were developed on the secondary, and up to five hundred in the primary. Mr. Fisher measured the voltages developed in both the secondary and primary audio-frequency transformers when the plate circuit is opened.

F. J. Somers presented a paper, "Characteristics of Electrostatic Loud Speaker." The paper described the construction of a type of electrostatic loud speaker and the method of testing its frequency response in comparison to a standard Jensen Dynamic Cone Speaker. Curves of response of both speakers and a comparison curve of the electrostatic speaker were shown. A model was exhibited.

Washington Section

On April 11th, in the Continental Hotel, Washington, D. C., the Washington section held their regular meeting. F. P. Guthrie, chairman of the section, presided. Eighty-two members and guests attended the meeting.

A paper, "Recent Developments in Radio in Application to Aircraft," by J. H. Dellinger and H. Diamond, was presented by H. Diamond. The paper discussed the radio aids to air navigation being developed and installed on the civil airways of the Department of Commerce. These aids are: communication, course navigation, and fog landing. The paper dealt chiefly with recent progress effected in the double-modulation type directive radio beacon. Methods are given for making this beacon serve four independent courses at variable angles to each other. A new triple-modulation beacon is described whereby twelve beacon courses are obtained making it possible to provide beacon service in every desired direction simultaneously. Special adaptations of the beacon system are described which give promise of facilitating landing in fog.

Messrs. Guthrie, Mirick, Robinson, Worrall, and Dorsey participated in the discussion which followed.

The election of officers was held with the following results: C. B. Jolliffe, chairman; T. McL. Davis, vice-chairman; John B. Brady, secretary-treasurer.

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GEOGRAPHICAL LOCATION OF MEMBERS ELECTED MAY 1, 1929

Elected to the Fellow Grade

England	London, SW7, 15 Rutland Mews	Slee, John A.

Transferred to the Member Grade

Arkansas	Conway, State Teachers' College	Cordrey, E. E.
California	Pasadena, California Institute of Technology	Mackeown, S. S.
Dist. of Columbia	Bellevue, Naval Research Laboratory	Dallin, Edwin B.
	Washington, Ouray Bldg	Brady, John B.
	Washington, 6310 5th St. N. W	Durkee, James D.
	Washington, 3100 Connecticut Ave., Apt. 105	Powell, Edwin L.
Illinois	Chicago, 1917 Warner Ave	Schoenwolf, F. L.
	Peoria, 147 Cooper Ave	Shalkhauser, Eric George
Massachusetts	Springfield, 162 Springfield St	Curtis, Leslie F.
New York	Schenectady, 9 N. Ferry St	Andrews, Paul D.
	Schenectady, River Road	Pike, O. W.

Elected to the Member Grade

Arkansas	Fayetteville, 143 S. Hill St	Barton, Loy E.
Dist. of Columbia	Washington, c/o Japanese Military Attaché, Portland Hotel	Kono, T.
Illinois	Chicago, 430 S. Green St	Frankel, Mortimer
New Jersey	Cranford, 31 Spruce St	Briggs, Loyd A.
New York	Brooklyn, 1244 East 35th St	Dreyer, Harry W.
England	Hull, c/o British Broadcasting Corp., 26 Bishop Lane	Calver, Frank N.
	London W1, Marconiphone Co., Ltd., 210 Totenham Court Road	Emery, E. J.
	Newcastle-on-Tyne, 5, Selborne Gardens, Jesmond	Nicholson, James K. A.
	Surrey, Windycot, Sugden Road, Long Ditton	Dryland, Alan G.
Straits Settlements	Singapore, c/o United Engineers, Ltd., Battery Road	Sutherland, J. G. A.

Elected to the Associate Grade

Arkansas	Little Rock, 224 Vernon St	Boone, Virgil
	Little Rock, 2510 West 16th St	Bracken, Ben
California	Berkeley, 1645 Capistrano Ave	Taylor, G. Wesley
	El Cajon, R. No. 1	Gruenwald, Harold C.
	Fullerton, 125 Ellis Place, Apt. 2	Hardison, Arthur A., Jr.
	Hemet, 311 E. Florida Ave	Sorkness, Hardus
	La Habra, 480 East First St	Inns, Stephen Henry
	Long Beach, 22 E. 4th St	Howell, Roger A.
	Los Angeles, 3176 Verdugo Road	Arnold, Harry
	Los Angeles, 3433 La Clede St	Borden, Charles
	Los Angeles, Western Air Express, 117 W. 9th St	Hoover, Herbert, Jr.
	Los Angeles, 1619 N. Benton Way	Stolle, R. A.
	Marshall, c/o R. C. A.	Case, Myron D.
	San Francisco, National Broadcasting Co., 111 Sutter St	Saxton, Alfred H.
	Santa Ana, 300 N. Broadway	Clark, William Moran
	Stanford University, Box 2287	Fisher, E. H.
	Venice, 650 California Ave	Gleason, Sterling
Connecticut	New Haven, 66 Norton St	Serig, Howard W.
Dist. of Columbia	Bellevue, Naval Research Laboratory	Bence, Clarence E.
	Bellevue, Naval Research Laboratory	Booth, H. A.
	Bellevue, Naval Research Laboratory	Ferree, John E.
	Bellevue, Naval Research Laboratory	Meyer, R. B.
	Bellevue, Naval Research Laboratory	Moore, Adam H.
	Bellevue, Naval Research Laboratory	Taylor, Samuel
	Bellevue, Naval Research Laboratory	Walker, Guy P.
	Washington, 502 Trenton St	Curtis, Westley F.
	Washington, 1821 Riggs Place N. W	Dent, Wm. U.
	Washington, 618 Morris St. N. E	Doremus, Cornelius W.
	Washington, 508 Rittenhouse St	Jett, Ewell K.
	Washington, 933 K St. N. W	Johnson, R. Stanley
	Washington, 2842 Vista St. N. E	Luke, E. L.
	Washington, 2316-32nd St. S. E	MacGregor, John J.
	Washington, 226-8th St. S. E	Noyes, Robert H.,
	Washington, 1901 K St. N. W	Stahl, Barton E.
	Washington, 1738 Mass. Ave. S. E	Talty, William B.
	Washington, 5125-8th St. N. W	Ullman, Mervin J.
	Washington, 1608-28th St. S. E	Wiseman, William W.
Florida	Miami, Pan American Airways, Inc	Sullinger, Ferris W.
Georgia	Atlanta, 585 Blvd. Place N. E. Apt. 8	Russell, B. A.
Illinois	Champaign, 804 South Prospect Ave	Duncan, Don C.
	Chicago, 1247 So. Kedvale Ave	Bank, Maurice

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Illinois (cont'd)	Chicago, 3030 N. Halstead St.	Fitzgerald, T. C.
	Chicago, 1145 Bryn Mawr Ave.	Lotter, John G.
	Chicago, 2416 Bryn Mawr Ave.	Moore, James S.
	Chicago, 216 N. Michigan Ave.	Paradise, Maurice E.
	Evanston, 1634 Chicago Ave.	Lund, Russell O.
	Pekin, Box 233	McGarvey, Harold R.
	Rockford, 130 No. Madison St.	Westlund, J. V.
	Urbana, 212 West Vermont St.	Hayden, Gilbert
	Wilmette, 530 Laurel Ave.	Shabino, Clarke L.
Indiana	Marion, c/o Y. M. C. A.	Detrick, Harold M.
	West Lafayette, 330 N. Grant St.	Brown, Wilfred E.
Kansas	Hamlin	Herold, Jos. L.
Kentucky	Louisville, Argonne Hotel	Tillett, Jesse
Maryland	Baltimore, 333 S. Gilmar St.	Ault, Arthur K.
	Baltimore, 2811 Chelsea Terrace	Krebs, William N.
	Brentwood, 4424 Charles St.	Dennary, James E.
	Mount Rainier, P. O. Box 86	Eisenhauer, Harry D.
	Mount Rainier, 3900–31st St.	Russell, Philip T.
Massachusetts	Boston, 7 Goodwin Place	Crawford, John D.
	Brockton, 93 Pleasant St.	Costello, John J.
	Cambridge, 362 Harvard St.	Stark, L. P.
	Holyoke, 49 Cleveland St.	Clemm, Robert Carl, Jr.
	Roxbury, 64 Brunswick St.	Salowitts, Louis W.
	Stoneham, 1 Franklin Road	Brown, Ronald W.
	Worcester, 45 Benefit St.	Gleason, Albert E.
Michigan	Ann Arbor, 715 Church St.	Byram, F. Cameron
	Battle Creek, c/o American Steam Pump Co.	Kinch, Oscar A.
	Detroit, 4217 McClellan Ave.	Carson, Walter E.
	Detroit, 1550 Elmhurst	Green, Walter M.
	Detroit, 1753 West Philadelphia Ave.	Sturman, George G.
	East Lansing, Box 611	Richards, A. R.
	Jackson, Spanks-Withington Co.	Bowmaster, J. M.
	Lakeside, Berrien County	Meier, H. G.
Missouri	Kearney	Thomason, Marvin
	St. Louis, 2551 W. Florissant Ave.	Cassens, Fred Benjamin
	St. Louis, c/o Moloney Elec. Co., 5390 Kingshighway N. W.	Wooley, W. C.
Montana	Bozeman, 301 North Grand St.	Woodhouse, Geoffrey Arthur
Nebraska	Falls City, 2012 Abbott St.	Quigley, James H.
New Jersey	Camden, 633 Linden St.	Heiser, Albert E.
	Hoboken, 406 Jefferson St.	Hohenstein, Jac.
	Newark, 149 Mt. Pleasant Ave.	Shea, Richard T.
	Newark, 218 Broad St.	Wasservogel, Gunther
	North Bergen, 1041 Bergen Line Ave.	Nickless, William H.
New York	Bayside, L. I., 201–14 50th Ave.	Lorch, Albert
	Brooklyn, 360 Bridge St., Room 712	Johnson, Richard M.
	Brooklyn, 220 Lincoln Place	Hammer, Robert H.
	Brooklyn, 87 Harrison Ave.	Seidman, Charles
	Brooklyn, 5509 13th Ave.	Seligsohn, Max
	Brooklyn, 683 Lincoln Place	Torrell, Roy Verner
	Croton-on-Hudson, Hamilton Ave.	Osborn, William R.
	Ellenville, 16 Hillcrest Ave.	Saylor, W. Edwin
	Elmhurst, L. I., 4910-94th St.	Kinsella, Christopher T.
	Garden City	Howe, Willard C.
	New York, 536 West 143rd St.	Donals, Alfred
	New York, 195 Broadway	Eller, Keith B.
	New York, 463 West St.	Koerner, Lawrence F.
	New York, 463 West St.	Penick, D. B.
	New York, 70 Van Cortlandt Park South	Volkmann, John E.
	New York, 371 Wadsworth Ave.	Welsh, Fred E.
	New York, WABC Columbia System, 113 West 57th St.	Whittaker, Edmund
	New York, 120 Broadway, Room 2235	Rosencranse, Charles R.
	Rochester, 6 Nothingham Circle	Bedford, Donald D.
	Schenectady, 129 South Country Club Drive, R. F. D. 1	Moles, Frank J.
Ohio	Cincinnati, 2535 Hemlock St. W. H.	Rouse, Louis
	Cleveland, 3033 Barber Ave.	Vasko, Albert James
	Cleveland Heights, 3927 Orchard Road	Glover, Gerald W.
	Columbus, 694 Carpenter St.	Storck, Howard C.
	Lakewood, 1299 Cove Ave.	La Ganke, H. C.
	Toledo, 1752 Pilgrim Road	Skeels, Dudley K.
Oklahoma	Tulsa, 1111 So. Denver St.	Jones, Burle R.
Oregon	Corvallis, Buxton Hall	Klein, William S.
Pennsylvania	Allentown, 617 No. St. Elmo St.	Dreisbach, Blair
	Allentown, 717 N. Ott St.	Peters, Arthur A.
	Allentown, Muhlenberg College	Wissler, Benjamin F.
	Bethlehem, 648 Highland Ave.	Bertolet, Benneville S.
	Catasauqua, 1122 N. 3rd St.	Urban, Paul Anthony
	Honesdale, 818 Court St.	Burlein, Lester F.
	Philadelphia, 3922 Pine St.	Lo, Tsu Zui
	Pittston, 105 So. Main St.	Ross, Stewart C.
	Slatington, 639 Main St.	Newhard, Stanley C.
Tennessee	Sparta	Haston, Claude William

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Texas	Fort Worth, P. O. Box 495	Noakes, G. A.
	San Antonio, Dodd Field, Fort Sam Houston	Marriner, Alfred W.
Virginia	Alexandria, 1006 Cameron St	Harrison, Edgar
	Cherrydale, 7 Mackey St	Jacobs, Edward N.
Washington	Spokane, E. 408 Sinto Ave	Dunnigan, Frank A.
	Spokane, 4611 N. Jefferson St	Storms, Henry J.
West Virginia	Cameron, Maple Avenue	Chambers, Albert V.
	Williamson, P. O. Box 323	Pirotte, Pete
Wisconsin	Milwaukee, 419–27th St	Nathness, Semore T.
	St. Croix Falls, Box 414	Powell, Wm. A.
Alaska	Ketchikan	Quist, Jesper S. M.
	Seward, Box 395	Dumarce, Herman
	Seward	Urie, Sol J.
Canada	Brandon, Manitoba, 627-13th St	Varcoe, H. R.
	Montreal, Que., c/o Northern Electric Co., 121 Shearer St	Gagnon, Elmore G.
	Montreal, Que., Canadian Nat. Ry., Radio Dept.	Olive, Gordon William
	Montreal, Que., Northern Electric Co., Ltd., Tech. Dept	Peachey, C. Arthur
	Montreal, Que., 7 Lafayette Ave., Montreal South	Platt, J. E.
	Montreal, Que., 3529 Bordeaux St	Walker, H. S.
	Montreal, Que., McGill University, Engineering Bldg	Wallace, G. A.
	Seaforth, Ont., Box 515	Daly, Edmund
	Toronto, Ont., 37 Lowther Ave	Bowers, E. J.
England	Yorks, Skipton, Spring Bank House	Slater, Wilfred
Japan	Sendai, No. 52, Kita-2-Bancho	Nagai, Kenzo
Scotland	Edinburgh, 11th Royal Crescent	Campbell, William

Elected to the Junior Grade

California	Merced, 861-20th St	Shaffer, Frank
	San Pedro, c/o Postmaster, USS *West Virginia*, USN	Macdonald, H. J.
Connecticut	Bridgeport, 165 Elm St	Harrington, Carlton E.
Michigan	Mt. Clemens, 175 Rathbone Ave	Wendler, Erwin, Jr.
Minnesota	Kenyon	Hjermstad, Hans U.
New Jersey	Trenton, 243 Warren St	Shaw, Elmer B.
New York	New York, 1064 Woodycrest Ave	Flythe, Thomas Y.
Ohio	Toledo, 1357 Elmwood Ave	Dupuis, Ernest F.
Pennsylvania	Coudersport, 730 N. Main St	Buchanan, Gurney Alfred

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APPLICATIONS FOR MEMBERSHIP

Applications for transfer or election to the various grades of membership have been received from the persons listed below, and have been approved by the Committee on Admissions. Members objecting to transfer or election of any of these applicants should communicate with the Secretary on or before June 29th.

For Election to the Associate Grade

California	Bolinas	Baker, Thomas S.
	Bolinas, Radio Corporation of America	Clark, Robert W.
	Bolinas	Riddle, Elmer R.
	Hollywood, 1719 N. Gardner St.	Blackburn, John F.
	Huntington Park, 6608½ Rugby St.	Hudson, T. L. W.
	Jackson, 115 Court St.	Snyder, W. Gilman, Jr.
	Los Angeles, 425 S. Bonnie Brae	Dixon, A. L.
	Los Angeles, 5533 Hollywood Blvd.	Wright, James L., Jr.
	Menlo Park, Box 163	Seal, H. Robert
	Palo Alto, Federal Telegraph Co.	Beatty, W. E.
	Palo Alto, 515 Oberlin Ave.	Lawrence, Robert B.
	Palo Alto, Federal Telegraph Co.	Suydam, C. H.
	Rosemead, 1223 S. Earl St.	Turner, W. E. H.
	San Pedro, USS *New Mexico*	Ruble, W. J.
Colorado	Denver, 2041 S. Josephine	Scheneman, Henry W.
Delaware	Wilmington, 418 S. Grant Ave.	Smyth, Earle
Dist. of Columbia	Washington, 621 G Street, S. W.	Appich, Wm. H.
	Washington, 3541 Holmead Pl., N. W.	Bradford, Henry K.
	Washington, 4637 Brewer Pl., N. W.	Burke, James W.
	Washington, Navy Department	Luckel, Frank
	Washington, Bureau of Standards	McIlwraith, Charles G.
	Washington, 1123 East Capitol St.	Simon, Isaac B.
	Washington, 1814 Jackson St.	Stevens, Waldo W.
Georgia	Columbus, Box 69	Kirkland, A. H.
	Columbus, Sta. WRBL	Shaffer, C. Robert
	Fort Benning, Headquarters Battery	Tate, Robert L.
Illinois	Chicago, 6429 S. May St.	Ellis, Robert M.
	Chicago, 1900 Mohawk St.	Fritzel, Joseph
	Chicago, 7614 Essex Ave.	Levy, Samuel L.
	Chicago, 1728 N. Luna Ave.	Owens, Robert F.
	Evanston, 139 Custer Ave.	Stein, Harold A.
Indiana	Anderson, 1219 Home Ave.	Miller, Louis
	Ashley	Urey, George M.
	Sweetser	Myers, Delbert
	Vernon, Box 24	Dausch, Elmer F.
Louisiana	New Orleans, 3111 Louisiana Ave. P'kway	Chateau, Arthur
	New Orleans, 512 St. Peter	Elkins, W. P.
	New Orleans, 1716 Painter St.	Peters, C. W.
Massachusetts	Cambridge, 7 Kirkland St.	Batchelder, L.
	Lexington, 16 Cliff Ave.	Chute, Dudley H.
	Lowell, 722 E. Merrimack St.	Coburn, E. D.
	New Bedford, 373 Pleasant St.	Meredith, W. B.
	Roxbury, 25 Fernboro St.	Katz, Louis S.
Michigan	Detroit, 9125 Avis Ave.	Anderson, Donald T.
	Detroit, 8881 Lauder Ave.	Jackson, Paul
	Detroit, 2923 Kirby W.	Kirby, Otto I.
	Detroit, 5842 Vernor Hgy West	Pennebaker, R. T.
	Lansing, 325 Smith Ave.	Wells, Lawrence V.
	Sault Ste Marie, SS *M. A. Bradley*	George, Edward M.
Minnesota	St. Paul, KSTP, St. Paul Hotel	Strout, Everett M.
Missouri	Kansas City, 3315 Holmes St.	Coates, B. Franklin
	Springfield, 906 East Grand	Moss, Terry L.
	St. Louis, 3328 Louisiana Ave.	Althoff, Frederick E.
	St. Louis, 5468 N. Kingshighway Blvd.	Cook, Jesse A.
	St. Louis, 3631 Bamberger Ave.	Proehl, Robert O.
Montana	Butte, 1120 West Gold St.	Walker, Raymond E.
New Jersey	Bloomfield, 18 Austin Place	Weaver, Karl S.
	Boonton, 302 Dawson Ave.	Cordell, Peter C.
	Collingswood, 106 Fern Ave.	Richards, Horace J.
	Deal, Box 122	Kerwien, Arthur E., Jr.
	Irvington, 41-42 St.	Coman, George E.
	Long Branch, 58 Washington St.	Emerson, Kenneth H.
	Merchantville, 606 W. Maple Ave.	Closson, Luke E.
	Newark, 820 Mt. Prospect Ave.	Canfield, Wilson R.
	Newmark, 180 Runyon St.	Dauber, Arthur O. F.
	Roselle, 133 Vine St.	Boeler, Gustave A.
	Trenton, Glen Cairn Arms Apts.	Miles, Lester F.

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New York	Brooklyn, 225—68th St	Marra, Anthony
	Brooklyn, 6 Bay 23rd St	Rosin, Arthur
	Brooklyn, 90 N. 9th St	Stromeyer, Charles F.
	Brooklyn, 60 Clinton Ave	Wood, Wilbert C.
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Part II

TECHNICAL PAPERS

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

RECIPROCITY IN ELECTROMAGNETIC, MECHANICAL, ACOUSTICAL, AND INTERCONNECTED SYSTEMS*

By

STUART BALLANTINE

(Radio Frequency Laboratories, Inc., Boonton, New Jersey)

Summary—*Recent criticism by Carson of the statement of the reciprocal relations in a radio communication system given by Sommerfeld is supported by a simple example showing the incorrectness of this statement.*

Carson's proof of the extention of Rayleigh's reciprocity theorem to a general electromagnetic system was limited to $\mu = 1$ and to sources consisting of ponderomotive forces on the electricity. A new proof is given under more general conditions, ϵ, μ, σ being merely restricted to be scalars and the impressed forces are of the electric type introduced by Heaviside and Abraham. These may be regarded as impressed charges and currents, including ether displacement current. The theorem is finally stated in terms of volume and surface integrals, and thus combines the viewpoints of both Lorentz and Carson.

The reciprocity relations in a mechanical system are reviewed.

The interconnection of electrical and mechanical systems is next considered and a "transduction coefficient" is defined. This concept is useful in formulating mechanical problems in electric-circuit form. An example of symmetrical transductance (copper coil in steady magnetic field) is given. The subject of units is taken up and it is proposed that in order to bring the mechanical quantities into agreement with the electrical ones when the latter are expressed in "practical" units, the mechanical quantities be expressed in "mechanical-volts," amperes, etc., i.e., in "practical-electric units of the mechanical quantities." A table for converting the principal mechanical quantities from c.g.s. to practical-electrical units is given.

Reciprocity is shown to exist in interconnected electro-mechanical systems with reversible transduction.

The equations of sound propagation in a gas are developed, regarding the velocity and excess pressure as the fundamental quantities. An acoustical reciprocity theorem involving these quantities is then proved.

Interconnections of mechanical and acoustical systems are then discussed and reciprocal relations are shown to exist in the composite system. Such relations are also valid for a system comprising electrical, mechanical and acoustical systems in series connection.

The reciprocity relations in a reversible electrophone are applied in a method of determining the frequency characteristic of an electrophone or microphone. This is called the "method of three electrophones." The overall transmission curves of the devices taken two at a time are measured and from these data the frequency characteristic of any one of them can be calculated. Only one of the electrophones is required to be reversible.

* Dewey decimal classification: R510. Original manuscript received by the Institute, February 26, 1929. Presented before joint meeting of the Institute and the International Union of Scientific Radiotelegraphy, American Section, at Fourth Annual Convention, Washington, D. C., May 15, 1929.

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AT this meeting there has been presented a criticism by Mr. J. R. Carson of the Sommerfeld-Pfrang reciprocity theorem for radio communication, which had been reviewed by me.[1] Mr. Carson seems entirely justified in emphasizing the superior generality and fundamental character of his generalization of Rayleigh's theorem, and in challenging the scope and validity of the Sommerfeld-Pfrang theorem in the form stated by Sommerfeld. The incorrectness of the Sommerfeld statement in terms of radiated power and received electric force in the case of an antenna which does not act as a dipol may be illustrated by a simple example, privately communicated by Mr. Lester Hochgraf.

Consider two equal antennas *A* and *B* some distance apart, oriented as shown in Fig. 1, and excited at such a frequency that there are complete positive and negative current loops. These antennas radiate equal power. Sommerfeld's theorem states that the "received field-strength" (*empfangene Feldstärke*) at *B* due to the radiation of a certain

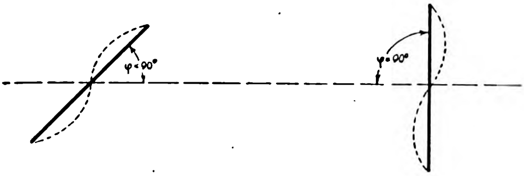

Fig. 1

average power from *A* will be equal to that at *A* with radiation of the same power from *B*. It will be seen, however, that the electric force produced at *B* by *A* is finite, whereas that produced at *A* by *B* is zero because *A* is in the equatorial plane of *B* and in this plane the radiation vanishes (at large distance).[2] This *reductio ad absurdum* seems sufficiently convincing and illustrates in a simple way the purport of Mr. Carson's criticism.

In passing it may be pointed out that the application of the theorem which was made in my review (loc. cit.) to the derivation of the transmitting properties of certain antennas over imperfect earth from their calculated receiving properties is in no way affected by this criticism. The only use to which the theorem was put there was to indicate the reciprocity existing between dipols or current elements, an application for which it is perfectly suitable. The theoretical conclusions of the

[1] Proc. I. R. E., 16, 513; April, 1928.
[2] Cf. Fig. 3, Ballantine: Proc. I. R. E., 12, 838; December, 1924.

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review concerning aircraft communication have since been verified experimentally.

2. Carson's Proof of Theorem II. The generalization of Rayleigh's theorem to extended electromagnetic systems given by Carson[3] (Theorem II) is of considerable importance. Carson points out that his proof is restricted to a medium in which $\mu = 1$. We may also note that his impressed forces are ponderomotive forces on the electricity and do not act to produce a displacement current in the ether.

The $\mu = 1$ restriction resulted from Carson's feeling that the current u should be a linear function of E. This does not appear to be necessary, however, and the restriction may be removed by adding to G a term $1/\lambda \cdot \text{curl } M$. The justification for this step follows from the fact that the added term, combined with u, vanishes when integrated over all space.

3. Proof of the Reciprocity Relations in an Electromagnetic System. An alternative proof of slightly more general reciprocity relations in an extended electromagnetic system may be constructed along the following lines.

In the first place all the constitutive quantities ϵ, μ, σ are allowed to differ from unity and may vary from point to point. In the second place the impressed forces will be regarded as purely electrical forces acting upon the total current; that is, upon the etheric displacement current as well as upon the current due to the motion of electricity. In other words the impressed forces will not be restricted, as in Carson's discussion, to ponderomotive or mechanical forces on the electricity. Generalized impressed forces of this type were included by Heaviside[4] and by Abraham.[5] A force acting only on the electricity can be replaced by an impressed current; a generalized force of the type now postulated, acting also on the ether, is equivalent to impressed electric and ether displacement currents combined with an impressed charge distribution. Impressed forces of the Heaviside-Abraham type may be objected to as artificial. It is true that in certain simple cases, as for example that of a generator winding revolving in a steady magnetic field, the seat of the impressed emf may be definitely localized, but in many other cases such forces are useful as a cloak for our ignorance of the exact mechanism of generation. In still other cases we may have sufficient knowledge of the generating mechanism but may wish to avoid unnecessary discussion by including only the

[3] J. R. Carson, *Bell System Tech. Jour.*, 3, 393, 1924.
[4] Heaviside, "Electromagnetic Theory," Vol. I, p. 36, (London).
[5] Abraham Foeppl, "Theorie der Electrizitaet," 6th ed., Vol. I, p. 216, (Berlin: 1921).

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terminals of the generator in the system. A shielded vacuum-tube generator would constitute an example of this. In such cases the idea of an impressed charge, with its attendant electric and ether displacement currents is perhaps more useful than the ponderomotive F forces acting directly upon the electricity.

Heaviside also introduced impressed magnetic forces and induction. The utility of these seems really questionable and they will be dispensed with here.

The medium is assumed to be linear and isotropic, that is, the constitutive quantities ϵ, μ and σ are scalars independent of the vectors, and the vectors which they connect are always in the same direction and linear vector functions of one another. No attempt will be made to prove that the reciprocity relations fail in the general case where these quantities are second-order tensors. Sommerfeld states, however, that in the case of propagation in a Faraday medium (e.g., ions in the earth's magnetic field) the reciprocal theorem is no longer valid. In this case the ϵ-tensor has the structure:

$$(\epsilon) = \begin{vmatrix} \epsilon_{11} & -i\alpha & 0 \\ i\alpha & \epsilon_{22} & 0 \\ 0 & 0 & \epsilon_{33} \end{vmatrix} \epsilon_{11} = \epsilon_{22} \neq \epsilon_{33}$$

which is skew-symmetric. Leaving aside intuitive suspicions, so far as I know it has not been definitely shown by Sommerfeld or anyone else that in this case, or in the case where (ϵ) is symmetric (i.e., $\epsilon_{jk} = \epsilon_{kj}$), or in the still simpler case when all components except ϵ_{11}, ϵ_{22}, ϵ_{33} are zero, the reciprocal relations fail or cannot be properly formulated.

Denoting the impressed electric force by E_0, the field equations for the quasi-stationary state (i.e., $\partial/\partial t = i\omega$) may be written in Gaussian units

$$E_0 = \frac{c}{4\pi\lambda} \text{ curl } \boldsymbol{H} - \boldsymbol{E} \tag{1}$$

$$\text{curl } \boldsymbol{E} = -i\omega\mu \boldsymbol{H}/c \tag{2}$$

$$\text{div } \mu \boldsymbol{H} = 0 \tag{3}$$

$$\text{div } \epsilon \boldsymbol{E} = 0 \tag{4}$$

$$\lambda = \sigma + i\omega\epsilon/4\pi. \tag{5}$$

Consider two independent distributions of impressed forces $\boldsymbol{E}_0'$, $\boldsymbol{E}_0''$, and corresponding total currents $\boldsymbol{C}'$ and $\boldsymbol{C}''$ $(\boldsymbol{C} = \lambda(\boldsymbol{E} + \boldsymbol{E}_0))$. Multiply (1) by $\boldsymbol{C} = c \text{ curl } \boldsymbol{H}/4\pi$ of the other distribution and subtract; then

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$$E_0'C''-E_0''C'=\frac{c}{4\pi}(E''\text{ curl }H'-E'\text{ curl }H''). \tag{6}$$

Remembering that

$$\begin{aligned} E''\text{ curl }H' &= \text{div }(H'\times E'')+H'\text{ curl }E'' \\ -E'\text{ curl }H'' &= -\text{div }(H''\times E')-H''\text{ curl }E'. \end{aligned} \tag{7}$$

Substituting values of curl E from (2) we have after some cancellation:

$$E_0'C''-E_0''C'=\frac{c}{4\pi}[\text{div }(E_1'\times H'')-\text{div }(E''\times H')]. \tag{8}$$

This is a point relation. Upon integrating over a region of space and using the divergence theorem, the result may be formally summarized in the following theorem.

Electromagnetic Reciprocity Theorem

If E_0' and E_0'' are two independent distributions of impressed electric force which act on the total current (convection, conduction, polarization, and displacement current in the ether), and C' and C'' represent the total currents $[=(\sigma+i\omega\epsilon/4\pi)(E+E_0)]$ resulting respectively from the action of E_0' and E_0''; if all quantities vary as $e^{i\omega t}$, and if the properties of the medium ϵ, μ, σ are scalars and independent of the field vectors, then:

$$\iiint [E_0'C''-E_0''C']dv=\frac{c}{4\pi}\iint [E'\times H''-E''\times H']_n ds \tag{9}$$

where the surface integral extends over the boundary of the region of the volume integration.

It may be inferred that the surface integral represents the effect of sources (i.e., E_0C) beyond it, so that if there are no sources outside the

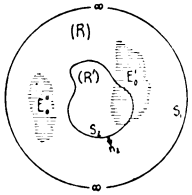

Fig. 2

region of integration the surface integral vanishes. For example, in Fig. 2 the region is shown bounded by a surface S_1 at infinity and a

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closed surface S_2. It is demonstrated in Note 1 that the S_1-integral vanishes. Hence

$$\iiint_R (\quad)dv = -\iint (\quad)_{n_2} ds \text{ (outward normal)} \tag{10}$$

but $\iiint_{R+R'}(\quad)dv=0$, hence $\iiint_{R'}(\quad)dv=\iint_{S_2}(\quad)_{n_2}ds$. This may be expressed in the following corollary:

Corollary I: *If under the conditions of the above theorem, the volume integration extends over all space and $\mu-1$, $\epsilon-1$, σ, and E_0 are finitely distributed, then*

$$\iiint_\infty [E_0'C''-E_0''C']dv = \frac{c}{4\pi}\iint_\infty [E'\times H''-E''\times H']_n ds = 0, \tag{11}$$

that is, the surface integral vanishes.

That this is true is almost evident by inspection when it is remembered that at a sufficient distance from a source the field vectors E and H become conjugate, that is, equal in magnitude and normal to each other and to the direction of propagation. To avoid interruption the detailed proof of this proposition is appended as Note 1 at the end of the paper.

The theorem in the form (9) involving both volume and surface integrals is more eloquent than the expressions of either Lorentz or Carson. Lorentz surrounded all sources with a surface integral; Carson used the volume integral over all space, assuming that all external influences could be accounted for by ponderomotive forces on the electricity. Theorem (9) combines both viewpoints. The surface integral is often useful in avoiding inquiries into complicated sources; it is simply necessary to surround the source by a surface and calculate $E\times H$ through it.

Equation (6) leads to another form. Since curl $H=4\pi C/c$, C being as before the total current;

$$E_0'C''-E_0''C'=E_0''C'-E'C'', \tag{12}$$

and

$$\iiint_\infty [E''C'-E'C'']dv=0. \tag{13}$$

This does not involve the impressed forces E_0 and is a general relation.

Considering for a moment the case contemplated in Carson's theorem, in which the impressed forces are of the nature of ponderomotive forces on the electricity and do not act on the displacement

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current (which is equivalent to saying that there are no impressed charges), equation (1) would be written

$$E_0=\frac{c}{4\pi}\operatorname{curl} H-\left(\lambda'+\frac{i\omega}{4\pi}\right)E \tag{14}$$

where $\lambda'=\sigma+i\omega(\epsilon-1)/c$. We should then have Carson's result (Theorem II)

$$\iiint_\infty [E_0'u''-E_0''u']dv=0. \tag{15}$$

Here u represents the current due to the motion of electricity. When we have to deal with impressed currents u_0 due to the motion of electricity (1) may be written

$$\operatorname{curl} H=4\pi\lambda E/c+4\pi u_0, \tag{16}$$

and the following reciprocal relations obtain

$$\iiint_\infty [E''u_0'-E'u_0'']dv=0 \tag{17}$$

note that at the field vector E replaces E_0 in this expression. An example of such an impressed current would be furnished by electrons acted upon by mechanical force.

If in (11) we substitute $C=\lambda(E+E_0)$ then:

$$\iiint_\infty [E_0'C''-E_0''C']dv=\iiint_\infty \lambda[E_0'E''-E_0''E']dv, \tag{18}$$

which furnishes another useful relation involving only the E's.

4. Example of the Application of the Electromagnetic Theorem. Calculation of the Effective Height of a Receiving Antenna. The physical significance of the reciprocity relations expressed by equation (11) may be brought out by means of a practical example. The following application to the calculation of the effective height of a receiving antenna is due to R. M. Wilmotte[6] and is simple and instructive.

The impressed voltage acting upon an element dx of the antenna (Fig. 3) is the component along the wire of the electric vector in the wave. Identify this with E_0 in (11) and call it $E(x)$ where x is the distance along the antenna wire. Our problem is to find the value of a localized emf E, which when inserted in some branch of the electric

[6] R. M. Wilmotte, *Jour.* I. E. E., 66, 966, 1928. Communication from the National Physical Laboratory, Teddington, England.

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circuit connected to the antenna terminals AB will produce in this branch the same current which is produced by the action of the distributed force from the wave. The reciprocity theorem states that in any element dx the current $I(x)$ produced by E will be equal to the

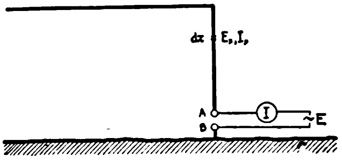

Fig. 3

current at the point of application of E due to an equal voltage $E(x)$ impressed at dx. Stated mathematically, for each element dx, $EI = E(x)I(x)$. This is true for each element of the antenna and is therefore true for the whole structure. Applying it to the whole antenna

$$E_0 I = \int_0^l E(x)I(x) \cdot dx. \tag{19}$$

We define the "effective height" as usual as $h = E/E'$; where E' is the electric force in the wave. If $E(x)$ is uniformly distributed along the antenna, $E(x)/E' = 1$. The possibility of this calculation rests upon knowing $I(x)$. Wilmotte,[7] in an important piece of experimental work, has verified in the case of an antenna of simple geometry, the approximate correctness of the old assumptions (Abraham, Pierce, etc.) of a sinusoidal distribution of current along the antenna.

In the case of a vertical antenna with E at the base, $I(x) = I \sin 2\pi x/\lambda / \sin 2\pi l/\lambda$, and

$$h = \frac{1}{\sin 2\pi l/\lambda} \int_0^l \sin \frac{2\pi x}{\lambda} \cdot dx = l \frac{4n_0}{\pi n} \frac{\sin^2 \pi n/4n_0}{\sin \pi n/2}, \tag{20}$$

where n_0 is the fundamental frequency and n the operating frequency. At the fundamental $n = n_0$ and $h = 2l/\pi$.

An interesting case is that of the short rod antenna proposed by F. H. Drake[8] for reception on airplanes in connection with a highly sensitive receiver. When the airplane is in horizontal flight the horizontal fusilage members contribute nothing, so that since $n = n_0$ the effective height is equal to half the actual height.

[7] R. M. Wilmotte, *Jour.* I. E. E., 66, 617, 1928.

[8] F. H. Drake, *Contributions from the Radio Frequency Laboratories* No. 8, October 1928; Proc. I. R. E., 17, 306; February, 1929; *Aero Digest*, October, 1928.

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5. Mechanical Systems. By a mechanical system is meant a system composed of moving parts having mass, compliance, and resistance (dissipation), these parts being usually amenable to direct physical observation. Elastic solids, fluids and gases are really also mechanical systems, but for convenience will be treated as acoustical. The reciprocal relations in a mechanical or dynamical system which can be described by means of linear equations have been discussed by the late Lord Rayleigh.[9] Reciprocity exists between the displacements and forces, although it is often convenient, especially when employing electric-circuit analogies of the mechanical system, to replace the displacements by the velocities.

If a set of impressed forces F_1', F_2', $F_3' \cdots$ produce displacements x_1', x_2', $x_3' \cdots$, and another independent set of forces F_1'', F_2'', $F_3'' \cdots$ produce displacements x_1'', x_2'', $x_3'' \cdots$, then

$$F_1'x_1''+F_2'x_2''+\cdots=F_1''x_1'+F_2''x_2'+\cdots, \tag{21}$$

or generally

$$\sum_n(F_n'x_n''-F_n''x_n')=0. \tag{22}$$

If we are dealing with velocities the x's in (22) are simply replaced by v's.

The circumstances of validity in the mechanical system are the same as those for an electrical system as discussed in Sect. 3, viz. quasi-stationary state with quantities varying as $e^{i\omega t}$, linearity in the relations between forces and displacements, and complete reversibility. Reversibility should not be confused with friction. Failure of reciprocity would follow if elastic hysteresis or an equivalent phenomenon were present. Friction will also introduce non-linearity when the velocity is small. All these practical limitations must be taken into consideration in making applications of the theoretical theorems.

6. Electro-Mechanical Systems. Since there are reciprocal relations between impressed electric forces and currents in an electric system, and between impressed forces and velocities in a mechanical system, they can be shown to hold in an interconnected electro-mechanical system, provided that the *interconnections* are linear and reversible.

In considering such composite systems I have found the use of an *interaction,* or *transduction coefficient* convenient. This may be defined as follows: let F be the mechanical force acting upon an element of a mechanical system as the result of a current I flowing in an element of

[9] Rayleigh, "Theory of Sound," Art. 108, (London: 1877).

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an electrical system. Then the *electro-mechanical transduction* T_{em} between these elements is defined as

$$F_0 = T_{em} I.$$

Similarly if E_0 represents the electromotive force due to an element of a mechanical system moving with velocity v, then the *mechano-electrical transduction* T_{me} is defined by

$$E_0 = T_{me} v.$$

If the transformation is reversible $T_{em} = T_{me}$; and if the interaction is linear T is independent of I or v. The dimensions of T are $[M^{1/2}L^{-3/2}]$. The "mutual energy" is equal to $1/2\ TIv$.

This concept is of considerable utility. Recent years have seen an increasing use of electrical analogues of mechanical systems. By this means we can bring to bear upon mechanical problems a great deal of knowledge concerning electrical networks which has resulted from the economic necessity for studies of electrical problems. In such formulations the role of the transduction coefficient is something like that of a mutual inductance between the real electrical circuit and the electrical analogue of the mechanical system.

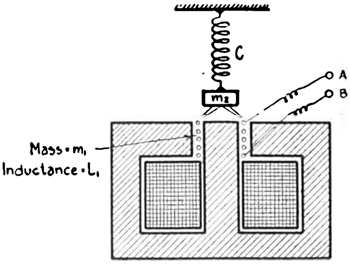

Fig. 4

As a simple example of this point of view consider the electromechanical system shown in Fig. 4. The coil is situated in a uniform magnetic field, H_0. We shall neglect eddy currents. The ponderomotive force in each element of volume is, generally,

$$F_0 = \frac{1}{c}[I \times B] - \frac{1}{8\pi} H^2 \operatorname{grad} \mu. \tag{23}$$

In the above case if the coil is of copper $B = H$, $\mu = 1$, so that

$$F_0 = \frac{1}{c}[I \times H_0] \tag{24}$$

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on each element of the coil. Hence $T_{em}=H_0/c$. In the mechano-electrical process the induced emf due to a motion of an element of the wire with velocity v is

$$E_0=\frac{1}{c}[v\times H_0] \tag{25}$$

so that $T_{me}=H_0/c$. The processes are thus reversible and the transductance is symmetrical, $T_{me}=T_{em}$.

The subject of units deserves some attention. If E_0 is expressed in volts, I in amperes, F in dynes, and v in cm per sec., then

$$\begin{aligned} F_0&=10^{-1}IH_0 \text{ dynes amp} \\ E_0&=10^{-8}vH_0 \text{ volts, cm/sec} \end{aligned} \tag{26}$$

therefore $T_{em}=10^{-1}$; $T_{me}=10^{-8}$. T_{me} can be adjusted to have the same numerical value as T_{em} when E and I are expressed in practical units by expressing F and v in appropriate units. We shall call these units in this particular case mechanical volts and mechanical amperes, respectively, or alternatively, practical electric units of mechanical force and velocity. A collection of the principal electro-mechanical analogues and of conversion factors from c.g.s. mechanical units to "practical electric" mechanical units is given in Table I.

TABLE I

Table for conversion of mechanical quantities expressed in c.g.s. units to units based upon the practical electrical system as may be required in the formulation of electrical analogues of mechanical systems.

Mechanical Quantity	Symbol	Definition	Electrical Analogue	c.g.s. Unit	Value in Practical Electrical units
displacement	x	x	charge	1 cm	10^{-8} mech. coulombs
force	F	F	voltage	1 dyne	10 mech. volts
velocity	v	dx/dt	current	1 cm/sec	10^{-8} mech. amps.
energy	W	Fx	energy	1 erg	10^{-7} joules.
power	P	Fv	power	1 erg/sec	10^{-7} watts
resistance	R	F/v	resistance	1 dyne/cm/sec	10^{9} mech. ohms.
mass	m	$F/i\omega v$	inductance	1 gram	10^{9} mech. henrys.
compliance	C	x/F	capacitance	1 cm/dyne	10^{-9} mech. farads
stiffness	s	F/x		1 dyne/cm	

According to this point of view the sytem in Fig. 4 may be represented as shown in Fig. 5. The transduction is represented by T, and symbolically as a mutual inductance. In this "mutual inductance" representation only the inductance directly concerned with the interaction is included in the primary coil and only the masses of the mechanical elements associated with the transfer are included in the secondary "coil." The so-called *motional impedance* of the system is easily seen to be the impedance between the electrical terminals AB, and the *damped impedance* that between AB with the secondary open.

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This representation in familiar electric form enables us to see at a glance what reactions the mechanical system will have upon the electrical system, and conversely the effect of the electrical circuit upon the motion of the mechanical system when an impressed mechanical force is acting. The latter effect is usually neglected by writers on this subject but obviously becomes of importance when the transfer is high. The conditions for maximum transfer between the systems can also be immediately formulated.

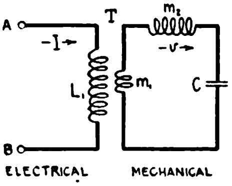

Fig. 5

The demonstration of the reciprocal relations in the interconnected system now proceeds along the usual lines. Let a set of impressed electric forces E_1', $E_2' \cdots E_n'$ act in the electrical system, and a set of mechanical forces F_1', F_2', $\cdots$, F_n' act in the mechanical system; let the corresponding currents in the electrical system be I_1', I_2', $\cdots$, I_n' and the velocities in the mechanical system be v_1', v_2', $\cdots$; let also F'', E'', I'', v'' represent a second set of such forces, currents and velocities. All quantities vary as $e^{i\omega t}$.

$$
\begin{aligned}
E_1 &= I_1 z_1 + I_2 m_{12} + I_3 m_{13} + \cdots v_1 T_{11} + v_2 T_{12} + \cdots \\
E_2 &= I_1 m_{12} + I_2 z_2 + I_3 m_{23} + \cdots v_1 T_{21} + v_2 T_{22} + \cdots \\
&\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots \\
F_1 &= I_1 T_{11}' + I_2 T_{12}' + I_3 T_{13}' + \cdots v_1 Z_1 + v_2 M_{12} + \cdots \\
F_2 &= I_1 T_{21}' + I_2 T_{22}' + I_3 T_{23}' + \cdots v_1 M_{21} + v_2 Z_2 + \cdots \\
&\cdots\cdots\cdots\cdots\cdots\cdots\cdots\cdots
\end{aligned}
\qquad (27)
$$

Here T_{jk} represents the electric force in mesh j due to velocity v_k of mechanical mesh k, and T_{jk}' represents the mechanical force in mechanical mesh j due to the electric current in mesh k; z and m represent electrical self and mutual impedances, and Z and M represent corresponding mechanical impedances.

Set up another system of equations for the other set of forces E'' and F''. Multiply line 1 of (27) by I_1'', line 2 by I_2'' and so on, and line 1 of the second set by I_1', etc., and subtract. Then if the connections

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are perfectly reversible, that is $M_{jk}=M_{kj}$; $m_{jk}=m_{kj}$, and $T_{jk}=T'_{jk}$, we obtain the following general reciprocal relations

$$\sum_n E_n'I_n''+F_n'v_n''=\sum_n E_n''I_n'+F_n''v_n'. \tag{28}$$

Considering the simple case in which all the forces except two are zero

$$E'I''=F''G', \tag{29}$$

which is an expression of simple reciprocity. In words this states that: if in an interconnected electro-mechanical system a unit electric force impressed in an electrical mesh produces a certain velocity in a mechanical mesh, then an equal mechanical force acting in the mechanical mesh will produce a current in the electrical mesh which is numerically the same as the velocity previously produced in the mechanical mesh.

Equation (28) is correct as it stands if F is expressed in dynes, v in cm per sec., and E and I are expressed in electromagnetic c.g.s. units, or electrostatic c.g.s. units. In order to be correct when E and I are expressed in volts and amperes, F must be expressed in *mechanical volts* and v in *mechanical amperes* as above. If E and I are in volts and amperes, and F and v are in dynes and cm per sec., then a factor of 10^7 will occur, that is

$$E'I''=10^7F''v'. \tag{30}$$

In the more general expression (28) this factor will also precede Fv whenever they occur.

A word may be said as to the linearity and reversibility in practical interconnections between electrical and mechanical systems. In the example of Fig. 4 reversibility may be safely assumed, but in general when the system contains ferro-magnetic materials, hysteresis and the non-linearity of the $B-H$ relation demand some caution. When the amplitudes are small the higher order effects can generally be ignored. Eddy currents not depending upon magnetic behavior will not disturb the reversibility or linearity.

7. Acoustical Systems. Turning now to the propagation of sound of small amplitude in air, it will be convenient for reference purposes to preface the discussion of the reciprocal relations by a brief statement of the principal equations. These will be given in modern vector form and the excess pressure p and velocity v will be regarded as the fundamental quantities, analogous to E and H of the electromagnetic field and F and v of the mechanical system.

If we ignore viscosity and have a ponderomotive force F_0 per unit mass acting upon the body of the gas, the motion may be described by Euler's equation

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$$F_0 = \frac{\partial v}{\partial t} + \frac{1}{2} \text{ grad } v^2 - v \times \text{ curl } v + \frac{1}{\rho} \text{ grad } P. \tag{31}$$

In addition we have the equation of continuity

$$\frac{\partial \rho}{\partial t} + \rho \text{ div } v + v \text{ grad } \rho = 0. \tag{32}$$

If the motion is irrotational, curl $v=0$, and if the amplitudes are small we may neglect the terms grad $(v \cdot v)$ in (31) and $v \cdot$grad ρ in (32); so that

$$F_0 = \partial v / \partial t + \text{grad } P/\rho; \tag{33}$$

$$\partial \rho / \partial t + \rho \text{ div } v = 0. \tag{34}$$

A further relation between ρ and P is furnished by considering the adiabatic expansion and compression of the gas

$$P/P_0 = (\rho/\rho_0)^\gamma, \tag{35}$$

where for air $\gamma = 1.4$, and P_0 and ρ_0 represent the steady values of these quantities. Now eliminate P from (31) and (32) by means of (35) and let the excess pressure $p = P - P_0$; then

$$F_0 = \partial v / \partial t + \text{grad } p/\rho_0, \tag{36}$$

$$\text{div } v = -\frac{1}{\gamma P_0} \frac{\partial p}{\partial t}. \tag{37}$$

These equations and the boundary conditions are sufficient to determine p and v. Take the divergence of (36)

$$\text{div } F_0 = \frac{\partial}{\partial t} \text{ div } v + \frac{1}{\rho_0} \nabla^2 p = -\frac{1}{\gamma P_0} \frac{\partial^2 p}{\partial t^2} + \frac{1}{\rho_0} \nabla^2 p_0, \tag{38}$$

from (37) so that p satisfies the wave-equation

$$\nabla^2 p - \frac{\rho_0}{\gamma P_0} \frac{\partial^2 p}{\partial t^2} = \rho_0 \text{ div } F_0. \tag{37}$$

Similarly, taking the gradient of (37)

$$\text{grad div } v = \nabla^2 v - \frac{1}{\gamma P_0} \frac{\partial}{\partial t} \text{ grad } p = \frac{\rho_0}{\gamma P_0} \left(\frac{\partial^2 v}{\partial t^2} - \frac{\partial F_0}{\partial \text{'}} \right), \tag{40}$$

so that v also satisfies the wave-equation

$$\nabla^2 v - \frac{\rho_0}{\gamma P_0} \frac{\partial^2 v}{\partial t^2} = -\frac{\rho_0}{\gamma P_0} \frac{\partial F_0}{\partial t}. \tag{41}$$

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The energy relations involving these quantities are interesting. Multiply (36) by $v\rho_0$ to get the rate at which work is done by the impressed force F_0 per unit volume:

$$\rho_0 F_0 v = \rho_0 v \partial v/\partial t + v \text{ grad } p. \tag{42}$$

But $v \cdot \text{grad } p = \text{div } (v \cdot p) - p \text{ div } v$. Substituting the value of div v from (37) and integrating over a region of space

$$\underset{(1)}{\iiint \rho_0 F_0 v dv} = \frac{\partial}{\partial t}\iiint\left[\underset{(2)}{\frac{1}{2}\rho_0 v^2} + \underset{(3)}{\frac{1}{2}\frac{1}{\gamma P_0}p^2}\right]dv + \underset{(4)}{\iint p \cdot v_n ds}. \tag{43}$$

This states that the rate at which work is done by the impressed forces (1) is equal to the rate of increase of the kinetic energy (2) plus the rate of increase of the energy of compression (3) plus the rate at which energy is escaping across the boundary (4). The flow of acoustic power is represented by the vector Pv, which is therefore analogous to the Poynting-vector $E \times H$ of the electromagnetic theory.

Reciprocal relations in acoustics have been stated by Helmholtz.[10] We shall, however, derive the relations anew and in somewhat greater generality in terms of the quantities F, p, and v which are here regarded as fundamental, rather than in terms of the "sources" and velocity-potential of the historical treatment.

Consider two independent sets of impressed forces F_0', F_0'', pressures p', p'' and velocities v', v''. Multiply (36) by the v of the other set and subtract

$$F_0'v'' - F_0''v' = \frac{1}{\rho_0}(v'' \text{ grad } p' - v' \text{ grad } p''), \tag{44}$$

in the quasi-stationary state (all quantities varying as $e^{i\omega t}$). But

$$v \text{ grad } p = \text{div } (vp) - p \text{ div } v;$$

hence, substituting the value of div v from (37),

$$\rho_0(F_0'v'' - F_0''v') = \text{div } (v''p' - v'p''). \tag{45}$$

This is a point relation. The result of integration over a region of space with the employment of the divergence theorem may be expressed in the following theorem.

[10] Lord Rayleigh, "Theory of Sound"; Vol. 2, p. 132, (London: 1878).

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Acoustical Reciprocity Theorem

If in an acoustical system comprising a medium of uniform density ρ_0 propagating irrotational vibrations of small amplitude, a distribution of impressed ponderomotive force F_0' per unit mass per unit volume varying as $e^{i\omega t}$ produces a velocity v' and pressure p', and a second distribution of impressed force F'' produces velocity v'' and pressure p'', then

$$\rho_0 \iiint (F_0'v'' - F''v')dv = \iint (v''p' - v'p'')_n ds, \qquad (46)$$

where the surface integral is to be taken over the boundaries of the region of volume integration.

Corollary I: *If the volume integral extends over all space and the distributions are finite:*

$$\iiint_\infty (F_0'v'' - F_0''v')dv = 0. \qquad (47)$$

The surface integral, as in the electrical theorem, represents the effect of impressed forces or other sources outside the region of volume integration, or those introduced at the boundaries. If there are no such sources the surface integral vanishes. This proposition is demonstrated in Note 2 at the end of the paper. Either the volume or the surface integral may be used, the surface integral being then taken over surfaces enclosing all sources.

The irrotational waves considered above are of a simple "compressional" type.

Body forces of the type F_0 postulated above are conceivable but are seldom encountered in practical acoustical problems. What we are generally concerned with are boundary forces arising from the motion of a mechanical boundary or discontinuity. The transfer of energy to the acoustical system may also be achieved by a thermodynamic process, as in the thermophone. An example of the F_0 type of body force would be the direct electrical excitation of an ionized gas; an example of the boundary type of force would be provided by a moving diaphragm, or piston. In this case the reciprocal relations are exhibited by the surface integral, that is

$$\iint (p'v'' - p''v')_n ds = 0. \qquad (48)$$

Either p or v may be regarded as "impressed." In the case of the radi-

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ation of sound from a vibrating piston, for example, either the impressed mechanical force on the piston or its velocity may be specified.

The above demonstration relates to sound propagation in a gas and should be supplemented by a similar demonstration for elastic solid and liquid media since the gas system may have such boundaries. The equations and the proof for irrotational vibrations in these media are the same as those given if the compressibility of the gas $(1/\gamma P_0)$ is replaced by the reciprocal of the bulk modulus $1/k$ for the solid or liquid, or $1/(\lambda+2\mu)$ for the solid.

It may be remarked in passing that a special type of force is required to set up pure rotational waves (transverse or shear waves) in an isotropic elastic solid, and in general this type of vibration will not be excited in the solid by an incident irrotational wave from the gas.

8. Mechano-Acoustical Systems. Let us now consider the reciprocal relations in interconnected mechanical and acoustical systems. A mechanically driven piston is an example of such interconnection. This sort of transduction differs from that between an electrical circuit and a mechanical system in that the mechanical and acoustic systems are really both mechanical. It is, in fact, more nearly analogous to the interconnection of an electrical network and an electrical space-

Fig. 6

radiation or receiving system, both of these processes being electrical. The forces in the acoustic system are mechanical forces, and the velocities are physical velocities, and both are measured in the same system of units which is used for the corresponding mechanical quantities. An extended argument is therefore not necessary to establish reciprocity in the connected system, and it is merely necessary to identify p and v of (46) with the force and velocity of the mechanical system. The acoustic pressure is the force per unit area of the mechanical system, and the velocities must be the same.

As an example of reciprocity in such a system consider two diaphragms A and B (Fig. 6). If we impress on a unit area at P_1 a force F_0' the velocity v' produced at a point P_2 of the other diaphragm will be equal to the velocity v'' which would be produced at P_1 if an equal force F'' were impressed on unit area at P_2. An important use of this theorem

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is the deduction of the receiving properties of a sound transducer from its transmitting properties, and vice versa.

9. Interconnected Electrical, Mechanical and Acoustical Systems. Reciprocal relations have been shown to hold in an electro-mechanical system and in a mechano-acoustical system, therefore they hold in a composite system involving all three systems in interconnection in

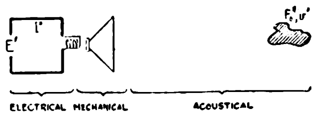

Fig. 7

this order. (Reversible transduction directly between electric and acoustical systems is rare and a mechanical system will usually intervene between them.) The reciprocity in such a composite electric-mechanic-acoustic system as shown in Fig. 7 may be expressed as follows, F_0 being an acoustical body-force

$$\iiint_\infty \rho_0 F_0'' v' w = E' I''. \tag{49}$$

If we deal with *pressure* in the sound field, p'' will replace $\rho_0 F_0''$. In words: if a generator of emf E' (in em or es c.g.s. units) operates in an electrophone circuit and produces at a point P in the sound field a velocity v', then a numerically equal force in dynes exerted at P on the gas material per unit volume, or communicated as a pressure, will

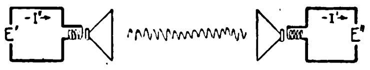

Fig. 8

produce a current I'' in the electrical mesh equal to the previously produced velocity v' at P.

The transduction must of course be reversible. An electromagnetic or electrostatic device (e.g., condenser microphone) would probably be sufficiently reversible, but a thermophone would certainly violate the requirements because of the irreversibility of the Joulean effect.

As a further extension of the composite system we may set up two reversible electrophones, as shown in Fig. 8, and state that in the terminating electrical circuits

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$$E'I'' = E''I'. \tag{50}$$

The reciprocal relations are then the same as if the system were entirely electrical.

10. Example of the Application of the Extended Reciprocity Relations. Frequency Calibration of an Electrophone. The above extended reciprocity relations between two reversible electrophones has an interesting and useful application in a method for determining the frequency response characteristics of an emitting electrophone or a receiving microphone. This may be called the "method of three electrophones" and is analogous to the method of measuring the resistance of a ground by the method of three grounds, or of the effective height of an antenna by using two additional antennas. The frequency characteristic of the electrophone usually desired is that in a certain direction, say normal to the sound-emitting area. (Characteristics in other directions may obviously be obtained in the same way.)

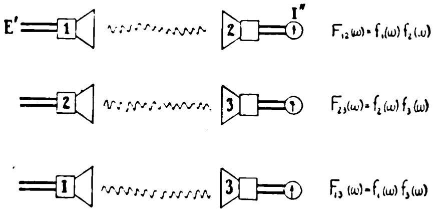

Fig. 9

The electrophones are used two at a time (Fig. 9) and are situated so far apart that the wave-front is substantially plane and the interaction between them is negligible. The purpose of this is to obtain data which will be fundamentally related to plane waves; investigations based on divergent waves may be carried out by proper adjustments of separation. One unit is excited by a generator E' of variable frequency and the current (or equivalent) in the other is measured. Let the frequency characteristics of the individual electrophones with respect to plane waves be denoted by $f_1(\omega)$, $f_2(\omega)$ and $f_3(\omega)$, respectively. Then for transmission from No. 1 to No. 2, for example, the measurement will give $F_{12}(\omega) = f_1(\omega)f_2(\omega)$. The transmission characteristic of No. 2 to No. 3 (F_{23}) is next obtained, and then that from No. 1 to No. 3 (F_{13}). This experimental information suffices to determine f_1, f_2, or f_3. For

$$\begin{aligned} F_{12}(\omega) &= f_1(\omega)f_2(\omega) \\ F_{23}(\omega) &= f_2(\omega)f_3(\omega) \\ F_{13}(\omega) &= f_1(\omega)f_3(\omega), \end{aligned} \tag{51}$$

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which is a set of simultaneous functional equations to be solved for the f's. The solutions are

$$\begin{aligned} f_1(\omega) &= (F_{12}F_{13}/F_{23})^{1/2} \\ f_2(\omega) &= (F_{12}F_{23}/F_{13})^{1/2} \\ f_3(\omega) &= (F_{13}F_{23}/F_{12})^{1/2}. \end{aligned} \tag{52}$$

If absolute values for the f's are desired care must be taken to measure properly the F's for each pair. For most purposes, however, all that is required is a curve of the relative variations of $f(\omega)$ with frequency; in this case absolute values of the F's are of no importance.

It may be noted that in the special case of two electrophones having identical frequency characteristics it is merely necessary to measure the frequency-transmission F for this pair since $F_{12}(\omega) = f_{12}(\omega)f_{21}(\omega)$ and $f_1 = (F_{12})^{1/2}$. This match will rarely be sufficiently exact in practical apparatus, except perhaps with precision microphones of the air-damped variety which have been accurately made and adjusted as to membrane tension.

It is also worth noticing that this method of calibration is equivalent to a Rayleigh-disk calibration in that it also takes into account the effect of reflection by the microphone.[11]

As will be seen from Fig. 9 this application of reciprocity requires only one of the electrophones (No. 2) to be *reversible*. Also note that one of them, No. 1, is always used as a source, and that the other, No. 3, is always used as a receiver. Thus a thermophone (irreversible) might be used at No. 1 as a source, with a reversible electrophone No. 2 as a comparator, to calibrate a carbon microphone, No. 3 (irreversible). It is needless to say that in order to obtain fundamental data the work should be performed in the open air and at a sufficient height to avoid disturbance due to reflection from the earth.

Note 1

It is required to evaluate the integral

$$\iiint_{\infty} [E_0'C'' - E_0''C']dv. \tag{1}$$

To do this we shall make use of the electromagnetic potentials A and ϕ, between which and the field vectors the following well-known relations exist

[11] For an account of this effect and of another method for evaluating it, see Stuart Ballantine, *Contributions from the Radio Frequency Laboratories*, No. 9, December 1928; *Phys. Rev.*, 32, 988, 1928; Proc. I. R. E., 16, 1639; December, 1928.

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$$B = \text{curl } A; \quad E = -\text{grad } \phi - i\omega A/c. \tag{2}$$

also
$$\Delta^2 A - \left(\frac{i\omega}{c}\right)^2 A = \text{dal } A = -i\omega E_0/c - 4\pi u/c - \text{curl } H \tag{3}$$

$$\text{dal } \phi = -4\pi\rho \tag{4}$$

$$\text{div } A = -i\omega\phi/c \tag{5}$$

$$M = (\mu - 1)H; \quad B = \mu H \tag{6}$$

u is the current due to the motion of electricity; ρ is the charge (exclusive of impressed charge, $\rho_0 = \text{div } E_0/4\pi$). Equation (6) of Sect. 3 may be written, since curl H = curl B − curl M:

$$E_0'C'' - E_0''C' = \frac{c}{4\pi}(E'' \text{ curl } B' - E' \text{ curl } B'' - E'' \text{ curl } M' + E' \text{ curl } M''). \tag{7}$$

Substituting for E and B their values (2) in terms of the potentials, we have

$$E_0'C'' - E_0''C' =$$

$$\begin{aligned}
&(1) \quad \text{grad } \phi'' \text{ dal } A' - \text{grad } \phi' \text{ dal } A'' \\
&(2) \quad + \frac{i\omega}{c}(A''\Delta^2 A' - A'\Delta^2 A'') \\
&(3) \quad + \text{grad } \phi'' \text{ curl } M' - \text{grad } \phi' \text{ curl } M'' \\
&(4) \quad + \frac{i\omega}{c}(A'' \text{ curl } M' - A' \text{ curl } M'').
\end{aligned} \tag{8}$$

We will consider these terms in order. Substitute in (1) the value of dal A from eq. (3), then

$$(1) \quad = -\text{grad } \phi'' \left(\frac{4\pi u'}{c} + \frac{i\omega E_0'}{c}\right) + \text{grad } \phi' \left(\frac{4\pi u'}{c} + \frac{i\omega E_0'}{c}\right) \tag{a}$$

$$-\text{grad } \phi'' \text{ curl } M' + \text{grad } \phi' \text{ curl } M''. \tag{b}$$

Note that term (1b) cancels term (3). Transform (a) by means of the theorem

$$\iiint D \text{ grad } \psi dv = \iiint \text{div } (\psi D) dv - \iiint \psi \text{ div } D \cdot dv$$

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and since div $C=0$

$$\text{(1a)} \quad = \frac{-i\omega}{c} \iiint (\varphi'' \, div \, \mathbf{E}' - \varphi' \, div \, \mathbf{E}'')dv$$

$$+ \frac{i}{c} \iint [\phi'(4\pi u'' + i\omega E_0'') - \phi''(4\pi u' + i\omega E_0')]_n ds.$$

If the integration embraces all space the integral over the infinitely distant surface vanishes for a finite distribution of E_0 and u. As to the volume integral, note that div $E = 4\pi\rho$, the value of which is given by (4) above. Making this substitution and applying Green's theorem

$$\text{(1a)} \quad = \frac{i\omega}{c} \iiint (\phi'' \Delta^2 \phi' - \phi' \Delta^2 \phi'')dv$$

$$= \frac{i\omega}{c} \iint \left(\phi'' \frac{\partial \phi'}{\partial n} - \phi' \frac{\partial \phi''}{\partial n} \right) ds.$$

If the surface is at infinity, $\partial/\partial n = -i\omega/c$ so that this term also vanishes. Terms (1)+(3) are therefore $=0$.

Term (2) vanishes for similar reasons. Applying Green's theorem

$$\text{(2)} \quad = \frac{i\omega}{c} \iiint (A'' \Delta^2 A' - A' \Delta^2 A'')dv$$

$$= \iint \left(\frac{\partial A'}{A \partial n} - A' \frac{\partial A'}{\partial n} \right) ds.$$

This can be regarded as a scalar equation for each component of A. As before, $\partial/\partial n = -i\omega/c$ for the infinitely distant surface so that the integral over all space is zero.

The remaining term (4) may be transformed by means of the theorem

$$\iiint P \text{ curl } Q dv = - \iint (P \times Q)_n ds$$

$$+ \iiint Q \text{ curl } P dv$$

so that

$$\text{(4)} \quad = \frac{i\omega}{c} \iiint (M' \text{ curl } A'' - M'' \text{ curl } A')dv$$

$$- \iint (A'' \times M' - A' \times M'')_n ds.$$

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If, as before, the distribution of $\mu-1$ is finite the surface integral vanishes. As to the volume integral, since curl $A=B$ and $M=(\mu-1)B/\mu$

$$(4)\qquad =\frac{i\omega}{c}\iiint\frac{\mu-1}{\mu}(B'B''-B''B')dv=0.$$

This completes the demonstration that the integral (1) vanishes when taken over all space.

Note 2

It is required to show that in the acoustical case

$$\rho_0\iiint(F_0'v''-F_0''v')dv=\iint(v''p'-v'p'')_n ds \qquad (1)$$

generally, and if the integration is over all space the result is equal to zero. Let there be two distributions of impressed force F_0' and F_0'' of the type $e^{i\omega t}$. Then according to equation (49)

$$\Delta^2 v'-k^2 v'=AF_0' \qquad (2)$$

$$\Delta^2 v''-k^2 v''=AF_0'' \qquad (3)$$

where $k^2=-\rho_0\omega^2/\gamma P_0$ and $A=-i\omega\rho_0/\gamma P_0$. Multiply (2) by v'' and (3) by v' and subtract

$$A(F_0'v''-F_0''v')=v''\Delta^2 v'-v'\Delta^2 v''. \qquad (4)$$

Integrate this over a region of space and transform the right-hand side by means of Green's theorem

$$A\iiint(v''\Delta^2 v'-v'\Delta^2 v'')dv=\iint\left(v''\frac{\partial v'}{\partial n}-v'\frac{\partial v''}{\partial n}\right)_n ds. \qquad (5)$$

If we consider an integral over all space and take the surface to be that of a sphere, $\partial/\partial n=-i\omega/c$, and the surface integral vanishes, giving

$$\iiint(F_0'v''-F_0''v')dv=0. \qquad (6)$$

The properties of the medium P_0 and ρ_0 have been assumed to be uniform. The theorem can also be proved for the case where they vary from point to point, but this paper is already long and this work will not be reproduced.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

RECIPROCAL THEOREMS IN RADIO COMMUNICATION *

By
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Summary—*Two reciprocal theorems, the generalized Rayleigh theorem and the Sommerfeld-Pfrang theorem are of great theoretical importance in radio-communication. A careful analysis of these theorems and their mathematical derivations shows that they are quite distinct and their practical fields of application different. In particular it shows that the Sommerfeld-Pfrang theorem labors under restrictions, implicit in its mathematical derivation, which seriously limit its field of practical applicability.*

In his "Theory of Sound", Rayleigh states and proves the following *reciprocal theorem:*

I. *Let there be two circuits of insulated wire A and B and in their neighborhood any combination of wire circuits or solid conductors in communication with condensers. A periodic electromotive force in the circuit A will give rise to the same current in B as would be excited in A if the electromotive force operated in B.*

This theorem is of the greatest usefulness in electric circuit theory and in communication engineering, and is employed extensively. One of its principal uses is to deduce the transmitting properties of a transmitter from its receiving properties, and vice versa.

In 1923 I was asked to look into the question as to whether the theorem is valid in radio-transmission. A brief examination showed that the proof of the theorem, as given by Rayleigh, is valid for quasi-stationary systems only, and is, in fact, deduced from the equations of rigid mechanics. It proved, however, not too difficult a matter to deduce the following generalized reciprocal theorem.[1]

IIa. *Let a distribution of impressed periodic electric intensity* $\mathbf{F}'e^{i\omega t}=\mathbf{F}'(x, y, z)e^{i\omega t}$ *produce a corresponding distribution of current density* $\mathbf{u}'e^{i\omega t}=\mathbf{u}'(x, y, z)e^{i\omega t}$, *and let a second distribution* $\mathbf{F}''e^{i\omega t}$ *produce a second distribution* $\mathbf{u}''e^{i\omega t}$ *of current density;* then

$$\int (\mathbf{F}'\cdot\mathbf{u}'')dv=\int (\mathbf{F}''\cdot\mathbf{u}')dv$$

* Dewey decimal classification: R190. Original manuscript received by the Institute, December 27, 1928. Presented before joint meeting of the Institute and the International Union of Scientific Radiotelegraphy, American Section, at Fourth Annual Convention, Washington, D. C., May 15, 1929.

[1] A Generalization of the Reciprocal Theorem, *Bell Sys. Tech. Jour.*, July, 1924.

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the volume integration being extended over all conducting and dielectric media. F and u are vectors and the expression $(F\cdot u)$ denotes their scalar product.

In 1925 Sommerfeld published the following reciprocal theorem[2]

III. *If A_1 and A_2 are two antennas located at O_1 and O_2 respectively, and have arbitrary orientations, and signals are first sent from A_1 and received by A_2 and then sent with the same average power from A_2 and received by A_1, then the intensity and phase of the electric field at the receiver A_1 will be equal to that previously produced at A_2, regardless of the electrical properties and geometry of the intervening media and the form of the antennas.*

In order to bring out the distinction between theorems IIa and III and to state the former in a form more suited for immediate application theorem IIa may be stated more concretely as follows:

II. *If an electromotive force is inserted in the transmitting branch of antenna A_1 and the current measured in the receiving branch of A_2, then an equal current (both as regards amplitude and phase) will be received in the transmitting branch of A_1 if the same electromotive force is inserted in the receiving branch of A_2.*

Just as theorem II is based on the more general theorem IIa, so the Sommerfeld-Pfrang theorem III is based on the following reciprocal theorem due to Lorentz.

IIIa. *If $E'e^{i\omega t}$, $H'e^{i\omega t}$ are the field vectors due to a periodic disturbance from a source A_1 located at O_1 and $E''e^{i\omega t}$, $H''e^{i\omega t}$ the corresponding field vectors due to a disturbance originating in A_2 from a source located at O_2, then*

$$\int_{1+2}[E''\cdot H']d\sigma=\int_{1+2}[E'\cdot H'']d\sigma$$

the surface integrals as indicated by the subscripts $1+2$ being taken over closed surfaces 1 and 2 surrounding the sources A_1 and A_2 respectively.

It now remains to discuss the utility and significance of these theorems in radio transmission problems and to point out the essential distinctions between them, concerning which there appears to be considerable confusion. Before doing so, however, it seems desirable, and even necessary, to discuss explicitly the restriction imposed on the

[2] *Zeit. für Hochfrequenztechnik*, 26, 93, 1925. In this paper Sommerfeld gives credit to H. Pfrang for the original statement of this theorem (München Dissentation, 1925) and at the same time points out that it is deduced from a reciprocal theorem due to Lorentz. The Sommerfeld paper is reviewed by Stuart Ballantine in the April 1928 issue of these PROCEEDINGS. In that review Ballantine mentions in passing that I have generalized Rayleigh's theorem for *electric circuits*. This would hardly seem an adequate description of the generality and scope of theorem IIa.

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generality of the theorems; restrictions which, as regards the Sommerfeld-Pfrang theorem III, seriously limit its applicability to radio transmission problems.

As regards the generalized Rayleigh reciprocal theorem, the only restriction is that the current (conduction plus polarization) must be linear in the electric intensity. Otherwise the medium may vary arbitrarily from point to point. In particular the transmitting and receiving antennas are absolutely unrestricted in their geometrical form, electrical constants, and disposition with respect to other conductors and conducting bodies, such as the earth. Both theorems fail when the waves are propagated in an ionized medium in which the earth's magnetic field has an appreciable effect on the conduction currents.[3]

In my original proof of the theorem the permeability μ of the medium was restricted to unity. Very recently Pleijel[4] has stated the theorem without this restriction. This theorem is not, however, the same as mine because the applied forces are not defined in exactly the same way and do not agree exactly with the physical definition of applied forces. The question as to whether the theorem holds when $\mu \neq 1$ seems to me still open.

In Sommerfeld's paper[5] and in Ballantine's review[5] thereof, certain restrictions on theorem III are carefully pointed out. For example, the relations $D=\epsilon E$, $B=\mu H$ and $I=\sigma E$, must be linear. On the other hand, other and in practice more serious restrictions are not stated but are implicit in Sommerfeld's mathematical proof. For this reason an analysis of his proof was made, which showed that the following restrictions limit the generality of the Sommerfeld-Pfrang reciprocal theorem III:

(1) The transmitting and receiving antennas, instead of being arbitrary geometrical forms, must both behave like simple electric or magnetic dipoles, as regards their radiation fields.

(2) The transmitting and receiving antennas must both be so far removed and isolated from other conducting bodies (*including the earth*) as to make the reflected field due to induced currents and charges in such bodies negligible in the neighborhood of the antenna which is regarded as the transmitter.

[3] This fact makes the application of the Reciprocal theorems to short wave transmission somewhat doubtful but does not seriously affect their application to long waves.

[4] Ingeniörs Vetenskaps Akademien, No. 68, 1927. There are two small errors in Pleijel's proof, which, however, are easily remedied.

[5] See footnote 2.

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These restrictions are so serious and so limit the applicability of the theorem in practical engineering problems as to call for the detailed analysis of the proof which follows:

Let $\boldsymbol{E}e^{i\omega t}$ and $\boldsymbol{H}e^{i\omega t}$ be the electric and magnetic vectors, respectively; $\boldsymbol{u}e^{i\omega t}$ the current density; and let it be assumed that the linear relations

$$\boldsymbol{D}=\epsilon\boldsymbol{E}; \qquad \boldsymbol{B}=\mu\boldsymbol{H}; \qquad \boldsymbol{u}=\sigma\boldsymbol{E} \tag{1}$$

hold everywhere in the medium.

Now suppose that $\boldsymbol{E}'$, $\boldsymbol{H}'$ denote the electric vectors due to a source A_1 located at point O_1; and $\boldsymbol{E}''$, $\boldsymbol{H}''$, the corresponding field vectors due to a source A_2, located at point O_2. The Lorentz reciprocal theorem then states that

$$\text{div}\ [\boldsymbol{E}'\cdot\boldsymbol{H}'']=\text{div}\ [\boldsymbol{E}''\cdot\boldsymbol{H}'] \tag{2}$$

the square brackets denoting vector product. It follows immediately that

$$\int_1 [\boldsymbol{E}'\cdot\boldsymbol{H}'']_n d\sigma_1+\int_2 [\boldsymbol{E}'\cdot\boldsymbol{H}'']_n d\sigma_2 = \int_1 [\boldsymbol{E}''\cdot\boldsymbol{H}']_n d\sigma_1+\int_2 [\boldsymbol{E}''\cdot\boldsymbol{H}']_n d\sigma_2 \tag{3}$$

the integrals 1 and 2 being taken over closed surfaces enclosing the sources A_1 and A_2, respectively. Subject to the linear relations (1), formulas (2) and (3), which I infer from Sommerfeld's paper are due to Lorentz, are perfectly general.

In applying these formulas, however, to deriving the reciprocal theorem III, the following approximations and restrictions are introduced:

(1) Over the surface of integration $\sum_1$ enclosing source A_1, the vectors $\boldsymbol{E}''$, $\boldsymbol{H}''$ are taken as *constant.* Similarly, over the surface of integration $\sum_2$ enclosing the source A_2, the vectors $\boldsymbol{E}'$, $\boldsymbol{H}'$ are taken as *constant.*

(2) In effecting the integration called for in formula (3), the vectors $\boldsymbol{E}'$, $\boldsymbol{H}'$ are assumed to be derivable from the Hertzian vector for either a simple electric or magnetic dipole; a corresponding restriction applies to the vectors $\boldsymbol{E}''$, $\boldsymbol{H}''$ on the surface $\sum_2$ enclosing the source A_2.

It is immediately evident that assumption (2) imposes very serious restrictions on the applicability of theorem III. These restrictions are that (1) the antennas A_1 and A_2 are simple linear oscillators, and (2)

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they are so far removed and isolated from other conducting bodies, (*including the earth*), that the secondary field due to currents and charges induced in such bodies is negligible in the neighborhood of the antenna which is regarded as the transmitter. In addition assumption (1) excludes the theorem from application to extended antenna arrays and the wave antenna. In view of these restrictions it seems to the writer that Sommerfeld's statement of the theorem is much too broad; in fact so broad as to be misleading.

Irrespective, however, of these restrictions, it seems to me that the generalized Rayleigh theorem II is of greater practical utility than the Sommerfeld-Pfrang theorem III. The former enables the engineer to deduce at once the transmitting properties of an antenna system from the receiving properties, and vice-versa. For example, the receiving properties of the wave antenna are quite simply deducible, to a good approximation, whereas a direct theoretical determination of its transmitting properties is extremely difficult. The generalized Rayleigh theorem furnishes an immediate answer to the problem.

On the other hand, in applying the Sommerfeld-Pfrang theorem, we require an adjustment of the two antennas for the same radiated power; a difficult matter whether by calculation or experiment. Furthermore in correlating the theorem with experimental results, the electric intensity is deducible only in terms of a detailed knowledge of the receiving properties of the antennas for reception from any direction or angle.

The generalized Rayleigh theorem says nothing explicitly about power and efficiency. For the very important case, however, where the impedances of transmitter and receiver are adjusted for maximum output amd maximum absorption of energy it can be shown that the ratio of the power output of the generator to the power absorbed by the receiver is the same for transmission in either direction. Thus, in the case of transmission between two entirely dissimilar antennas the transmission efficiency is the same in the two directions provided the terminal impedances are properly adjusted. This proposition is not immediately obvious but is very simply deducible from the Rayleigh theorem.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

THE MEASUREMENT OF DIRECT INTERELECTRODE CAPACITANCE OF VACUUM TUBES*

By
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Summary—*A method of measuring direct capacitance in the range* 10^{-10} *to* 10^{-15} *farads by a charging current at radio frequencies is described. A low resistance current indicator is shunted across one of the direct capacitances. The charging current* $I=\omega EC$ *is a measure of the direct capacitance provided that* ωE *is maintained constant. A simple circuit arrangement provides an apparatus giving reliable results measured visually by substitution with a standard.*

The method has been practically applied in the measurement of direct capacitances present in vacuum tubes providing data for design purposes and for control of product uniformity. The method is specially advantageous in its ability to measure exceedingly small direct capacitances present in screen-grid tubes.

With the development of more efficient electric circuits employing the vacuum tube, the determination of the magnitude of the individual or *direct* interelectrode capacitances became of engineering importance. The input impedance of a vacuum tube is calculable only when the direct capacitances, the "constants" of the tube, and the associated circuit dimensions are known.[1]

A three-electrode tube forms a simple network of three direct capacitances. Where there are n electrodes, the number of direct capacitances N is

$$N=\frac{n}{2}(n-1)$$

Maxwell,[2] in determining the direct capacitances present in a guard ring capacitor, used an absolute charging current method and an electrometer indicator. More recently, in a general discussion of direct impedances, Campbell,[3] has described a simple ammeter method. This method has been used by Hull[4] for the measurement of the extremely small direct capacitances present in screen-grid tubes. With modifications, it forms the basis of the method used by the present authors.

In order to obtain data for purposes of tube design and for purposes

* Dewey decimal classification: R220. Original manuscript received by the Institute, March 6, 1929.

[1] Nichols, *Phys. Rev.*, 13, 404–419, 1919. Miller, Bul. Bur. Std. Scientific Paper No. 351. Hartshorn, *Proc.* Phys. Soc., 39, 108, 1926.

[2] Maxwell, *Electricity and Magnetism*, 1, 350, ed. 1892.

[3] Campbell, *Bell Sys. Tech. Jour.*, 1, 18–38, 1922.

[4] Hull, *Phys. Rev.*, 27, 437, 1926.

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of quality control of product uniformity, it is necessary to measure direct capacitance with a high degree of precision in a reasonably rapid manner. In consideration of the anomalous frequency behavior of the dielectric of the tube and the high resistance metal film, called "getter," on the inside walls of the glass bulb, and in consideration of leakances of a very high order but comparable with the capacitive reactance of the small direct interelectrode capacitances at audio frequencies, the optimum of accuracy of capacitance measurement of receiving tubes was found to occur with measurements obtained by a charging current method at about one-half megacycle. Practically, the effective direct capacitance at radio frequencies is principally desired although it is about five per cent lower than the capacitance measured at audio frequencies due to the above mentioned effects.

In complex capacitance networks, as exemplified by the screen-grid tube, the errors of the Schering direct capacitance bridge or the Campbell-Colpitts bridge may exceed the capacitance sought. For information regarding the Schering bridge see the reference to Hartshorn.

The charging current method is flexible in permitting extremely small capacitances in the range 10^{-10} to 10^{-15} farads to be measured with known accuracy. The method can be made simple and visual and provides easily obtained results that are reliable.

The Theory of the Method

The fundamental circuit arrangement for the charging current method is shown in Fig. 1.

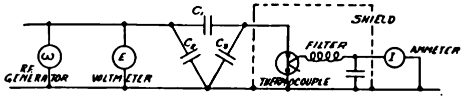

Fig. 1—Showing the Fundamental Circuit Arrangement for the Charging Current Method for Measuring Small Direct Capacitance.

The high-frequency generator supplies voltage sufficient to cause a measurable current to flow in the branch C_1 and the thermocouple. This current is not affected by C_2 if the voltage is held constant, nor by the shunting effect of C_3 and the capacitance of the leads to the shield since the reactance of these capacitances is high compared with a low resistance thermocouple. The error introduced by the shunting effect is easily calculable and by choice of thermocouple resistance this error may be made vanishingly small for capacitance dimensions encountered in vacuum tubes up to the 100-kw size. In order to have

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the current indicator at ground potential it is connected to the thermocouple through a filter in the manner shown.

It is evident that the precision of the measurement depends on the accuracy of the determination of the three quantities—voltage, current, and frequency. To measure the three quantities independently would require excessively refined technique and apparatus to obtain precise results. However, in practically applying the fundamental method a simple solution has been developed.

The Practical Circuit Arrangement

The application of this fundamental method requires some alteration of circuit arrangement to make it direct reading. Fig. 2 shows the circuit.

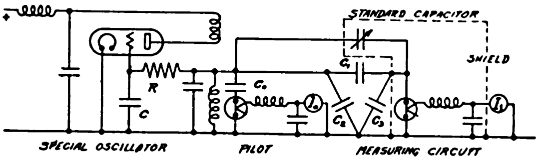

Fig. 2—Showing the Practical Circuit Arrangement for the Charging Current Method for Measuring Direct Capacitance of Receiving Tubes.

In the above figure the network at the left represents an oscillator which is so adjusted that when the frequency is lowered by the addition of capacitance across the oscillating circuit, the terminal voltage rises sufficiently to maintain constant the product ωE. This constancy is indicated by a pilot consisting of a very small fixed capacitor C_o and an indicator of its charging current I_o.

The current charging a capacitance is

$$I = \omega E C$$

Where ω = angular velocity of fundamental frequency

$$E = 0.707\sqrt{E^2_{1m} + 4E^2_{2m} + 9E^2_{3m} + \ldots}$$

in which series E_{1m}, E_{2m}, E_{3m}, etc., express the maximum value of the voltage of the fundamental frequency, the second harmonic, the third harmonic, etc., respectively. A square law symmetrical indicator of current, like the vacuum thermocouple used, obeys the law of superposition of vector quantities, very exactly in this case where the harmonics present are small compared with the fundamental.

In the above equation, if C_o is constant and the current indicator, I_o, shows a constant deflection, obviously ωE must be constant also. The use of a special oscillator is obligatory only when special accuracy is required in rapid measurement as otherwise ωE may be held con-

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stant manually. The method thus escapes from the difficulties encountered in measuring frequency and voltage independently.

A likewise simple use of a standard substitution capacitor eliminates the need of measuring the absolute value of the current and thus there is no need for calibration of the current indicator. The constancy of

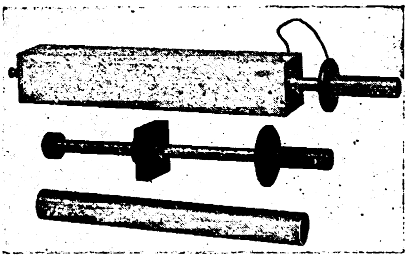

Fig. 3—Showing a Standard Capacitor Having a Capacitance Increment of 2.00 $\times 10^{-12}$ farad per inch of cylinder length.

the calibration of the standard capacitor is all that is required and the use of coaxial cylinders makes this easily possible.

Method of Measuring Three-Electrode Receiving Tubes

With the standard capacitor set at any convenient arbitrary zero, the tube, represented in the figure by C_1, C_2, and C_3, is inserted in a

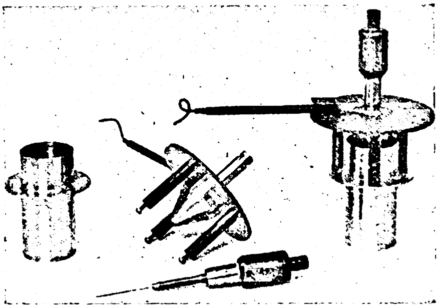

Fig. 4—Showing a Standard Capacitor Having a Capacitance Increment of 1.00×10^{-14} farad per turn of the micrometer head.

shielded socket fixture and the reading of the meter I_1 is noted. On removal of the tube, the standard capacitor is increased until I_1 reads the same as noted. Then the increment of the standard capacitor is equal to the direct capacitance of the tube, C_1 in the figure. This is true only, if meanwhile, the pilot circuit indicator I_o has shown no change in deflection.

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It is feasible to provide a circuit arrangement which will allow the meter, I_1, to be scaled off in micromicrofarads, thus providing a "direct capacitance meter."

The Standard Capacitors

The standard substitution capacitors are of the calculable increment coaxial cylinder type calibrated linearly in micromicrofarads. Physicists have used this type of capacitor for many years. The calculation is given in Bul. Bur. Std. Circular No. 74, p. 236. An interesting work on the determination of errors in coaxial cylinder standards has been published.[5]

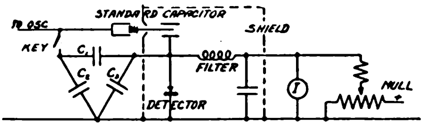

Fig. 5—Showing a Sensitive Current Indicator and Small Increment Standard Capacitor Arranged for Measurement of 0.0005 to 0.100 μμf of direct capacitance.

Fig. 3 shows the photograph of a standard capacitor most generally used in vacuum-tube measurements. It consists of three parts, a movable inner cylinder having a well insulated handle and connected to the oscillator by a flexible lead, a fixed outer cylinder connected to the thermocouple, and a grounded outer shield which also provides mechanical support for the two insulated cylinders.

Fig. 4 shows a photograph of a similar standard capacitor having a capacitance increment of 1.00×10^{-14} farad per turn of the micrometer head. It is especially used to measure the extremely small feed-back capacitance of the screen-grid type of tube.

Special Measurement Technique for Screen-Grid Tubes

The measurement of the screen-grid tube feed-back capacitance C_{pg} requires a more sensitive current indicator than the thermocouple. A carborundum fixed detector is used to indicate the small charging current. A filtered balanced circuit permits the use of a microammeter and provides a sensitivity of about 1 μa per 1/100 μμf of direct capacitance. Due to the resistance of the detector, the shunting effect of the direct capacitances is not negligible, and the tube is not removed from its measuring position but the oscillator lead is broken instead when the standard substitution is made.

The network shown in Fig. 5 represents the circuit used for the measurement of extremely small direct capacitances.

[5] M. Bedeau, *L'Onde Electrique*, 5, 613, 1926.

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As has been pointed out by Moullin[6], an asymmetrical non-linear circuit element, like a crystal detector, is not a rigorous indicator of the current flowing in the circuit, as it is affected by wave form and does not indicate true effective current. However, the crystal has such an advantage in simplicity over any equally sensitive instrument whose indication is independent of wave form that it is used here in spite of its somewhat lesser accuracy. For this reason it is preferable in accurate measuring practice to employ the measuring circuit itself (Fig. 5) as the pilot in the following manner. Without the tube in the measuring position, let the residual capacitance of the standard capacitor and

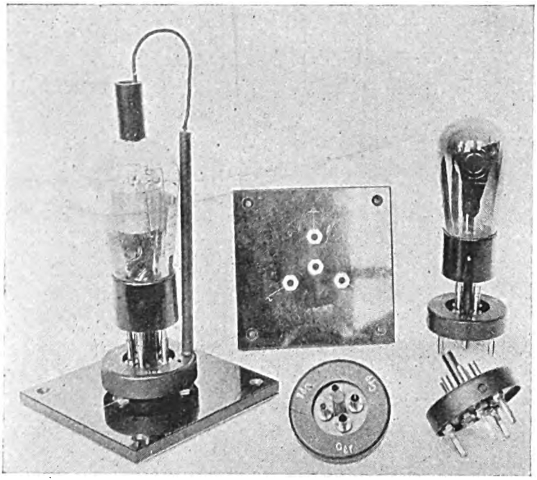

Fig. 8—Showing the Shielded Fixture Used for Measuring C_{gp}, C_{gf}, and C_{pf} by a Simple Rotation of the Fixture on the Instrument Panel.

the crystal detector serve the same function as C_o and I_o in Fig. 2. If connecting the tube to be measured across the oscillator terminals results in no change in the current indicator I in Fig. 5, then it is evident that the oscillator is properly adjusted and the measurement may proceed.

Power Tube Capacitance Measurements

The general class of power tubes are measured on a large metal sheet which is the ground plane. Two shielded flexible low capacitance cables make the necessary oscillator and thermocouple connections to the elements of the tube being measured. Heavy flexible ground connection is made to the elements that are grounded and care is

[6] Moullin, *Jour.* I. E. E., 298, 1926.

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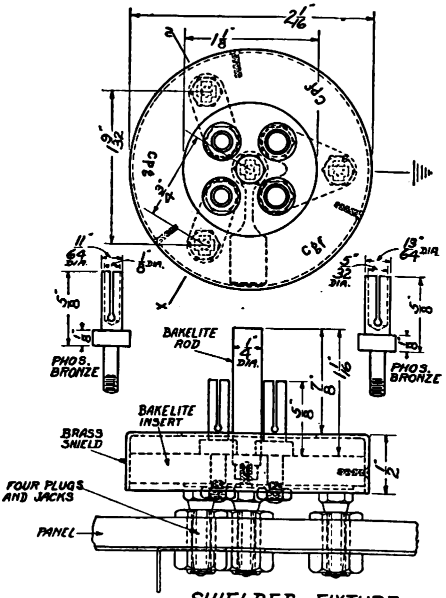
taken to provide contacts to the elements that are of negligibly small resistance. A low resistance shunt is provided across the thermocouple of the measuring circuit for limiting the thermocouple current when measuring large direct capacitance. In other respects the measurement procedure is the same as for capacitances of smaller dimensions. Fig. 7 shows a photograph of a power tube on the apparatus used for general direct capacitance measurements.



SHIELDED FIXTURE
(SAME WITH UY PIN SPACING)

Fig. 9

Fixtures Used

The measurement of the direct capacitance of receiving tubes can be made independent of the leads to the tube and the external disposition of the wiring only by the use of a special shielded standardizable fixture. Ordinary sockets are out of the question due to expanding prongs and errors involving 0.1 $\mu\mu f$ in stray coupling capacitance to parts of the tube being measured. A fixture devised to insure independence of the geometry of the leads and to insure reproducibility of the measurement is shown in Figs. 8 and 9. A simple rotation of the shielded fixture on the instrument panel permits the measurement of C_{pg}, C_{gf}, C_{pf}. The UX and UY types of fixtures differ only in their pin spacing.

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Apparatus for Receiving Tube Capacitance Measurements

The ensemble of apparatus used for receiving tube measurements is shown in Fig. 10. One side of the apparatus is provided for the measurement of feed-back capacitance of the screen-grid type of tube

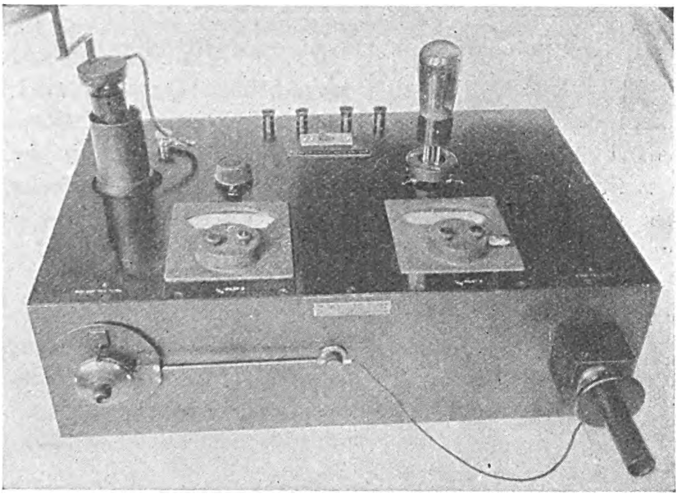

Fig. 10—Showing the Ensemble of Apparatus Used for Receiving Tube Capacitance Measurements.

and the other side for the measurement of the direct interelectrode capacitance of the ordinary standard type of receiving tube. The figure shows the handles of the standard capacitors which are operated from the front of the metal case and shows two tubes mounted in the measuring position, one a screen-grid tube in a shield cup and the other a three-electrode tube in the shielded socket fixture previously described.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

PHOTORADIO DEVELOPMENTS*

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General Picture Developments, 1926-1928

OUTSTANDING in this period has been the work of the American Telephone and Telegraph Companyin the extension of their commercial picture network throughout the United States. A fine example of the work they are able to perform is given herewith. (Fig. 1) It is interesting to note that the telephone service has been used extensively for the transmission of printed matter—bond circulars in particular.

The other wire line method that continues to be used commercially is the Bartlane System handled between the *Daily Mirror* in London, and the Pacific and Atlantic Photo Service in this country. The method consists in reducing a picture to a perforated tape of the same sort as is used for cable messages, and then this tape is transmitted over the Western Union System across the water. The picture is then reformed from the holes punched in a duplicate tape on the other side of the Atlantic.

On December 1st, 1927, wire picture service was inaugurated between Berlin and Vienna by the Karolus-Telefunken System. Two novel features of this system must be mentioned. The first is the use of the Kerr cell. The Kerr effect has been hidden in the archives of science for forty years, and it remained for Dr. Karolus to bring it to practical use in the transmission of pictures. The Kerr cell consists of a gap between two oppositely charged electrodes with a solution of nitrobenzol covering them. Such a combination has the ability to change the speed of light passing through the cell depending upon the plane of polarization of that light. Specifically, if the light is vertically polarized with respect to the surface of the electrodes, the light train will pass through quicker when potential is applied between the electrodes; if the light is in the same plane as the electrodes it will be retarded.

To make use of this phenomenon, a nicol prism is turned to give a beam of light at 45 deg. with the electrode surfaces on entering the

* Dewey decimal classification: R582. Original manuscript received by the Institute, March 6, 1929. Presented before New York section of the Institute, January 10, 1928; before Los Angeles section, August 27, 1928; before San Francisco section, September 4, 1928.

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cell. A second nicol prism is then used on the side from which the light leaves the cell, but it is turned to allow only light at 90 deg. with respect to the first prism to pass. The effect of this optical system is to allow no light to pass when there is no electrostatic distortion produced in the cell. Now, however, when voltage is applied to the electrodes, the distortion will take place. A vector diagram shows that if the horizontal component is slowed down half a wavelength with respect to the vertical in passing through the cell, the emerging light will be plane polarized at 90 deg. from the direction in which it entered, and will therefore then be able to pass through the second or analyzing

(original)

(received)

Fig. 1

nicol prism perfectly. This means that with the right amount of voltage applied to the electrodes, we will get a change from no light to full light, and the outfit therefore works as a simple light valve. It has the great advantage of being practically inertialess. The practical voltage range for operation is of the order of 500 to 1000 v. The slight disadvantages of the method are due to the fact that nitrobenzol is not a perfect dielectric and allows some current to flow, which causes some deterioration, and secondly the response is linear only over a restricted part of the curve, but its speed action is a tremendous asset.

The other outstanding development of this system is the photocell construction for which Dr. Schriever of the Telefunken Company is responsible. Instead of being in the more usual Christmas tree ball

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form, this photocell is much like a very flat doughnut. (Fig. 2) The purpose is two-fold; first to simplify the light system, and second and most important, to increase the available light change in the pick-up at the transmitter some fifty-fold.

To accomplish this, at the picture transmitter, this doughnut cell is placed directly in front of the picture transmitting cylinder. A very intense beam of light is then focused down to a very fine point through the hole of the doughnut at the picture. This fine little spot of light which seems to be a little glowing fire then reflects back on to the solid part of the glass photocell, and acts on the light sensitive

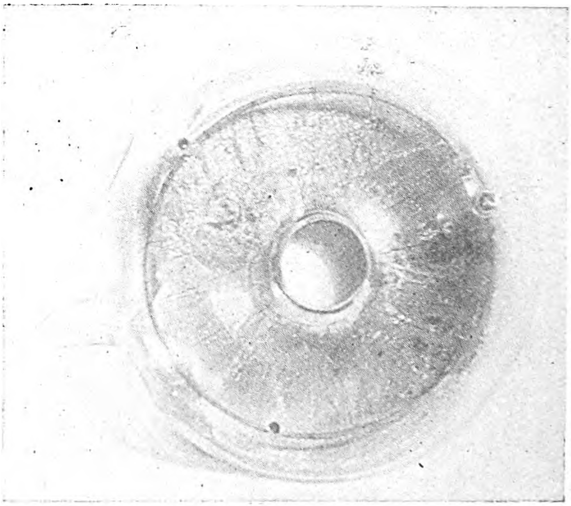

Fig. 2

chemicals placed there. Due to the fact that the active part includes such a large angle from the little spot of light, practically all the available light is used. In this sense it corresponds to having a lense aperture of FO.5, and all camera enthusiasts know what that would do in the way of speed and intensity of action.

Another entrant into the picture transmission work is the Marconi Company, with Mr. Wright heading their activities in this direction. He is working on the principle of reversing the usual method of analysis at the transmitter and receiver—he uses a stationary cylinder at each end. The picture is slid forward in a curved form along the length of the cylinder. Half way around the cylinder there is a fine slit corresponding to a line of the picture, and as the picture is moved forward, a fine light spot is rapidly revolved inside the cylinder to cross the pic-

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ture from one side to the other, one line at a time through this slit, as the picture advances. He is working at quite high speeds, planning to take full advantage of the higher speed capabilities of the Marconi radio beam.

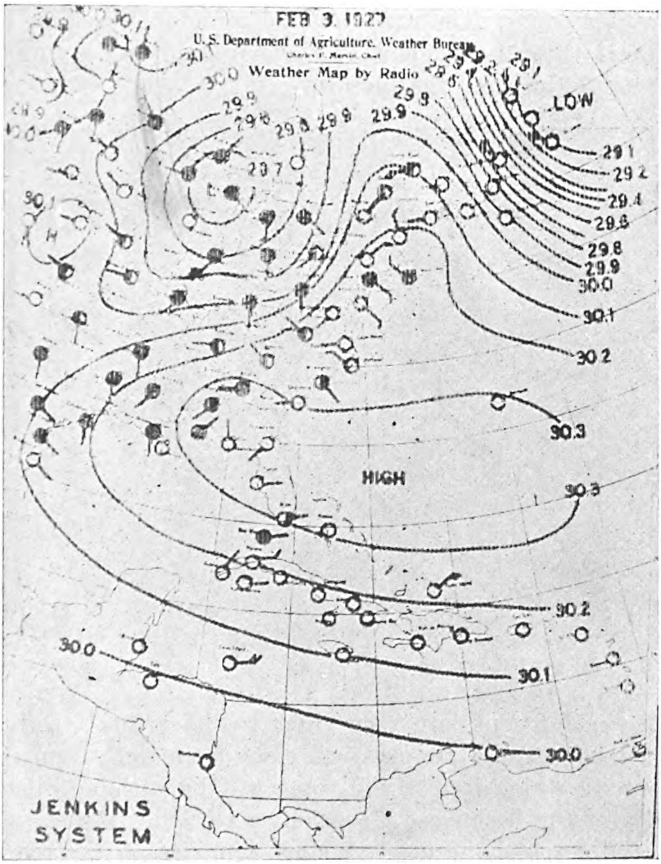

Fig. 3

Map Work

Two of the regular workers in picture radio, C. Francis Jenkins of Washington, and Dr. Max Dieckmann of Munich, Germany, have modified their equipment to simplify the recording of weather maps at sea. (See Figs. 3 and 4, respectively.)

Amateur Reception

Several workers have tried to interest the amateur in picture reception. They are—T. Thorne-Baker of London, one of the long

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standing workers in the field, Mr. Jenkins, and, more recently, Austin G. Cooley, in connection with *Radio Broadcast* of Garden City, Long Island. Mr. Cooley makes use of the old phonograph that may have been otherwise discarded in the amateur's home, and specializes in a "corona" discharge at the end of a fine needle point resting on photographic paper. The variations in the intensity of the incoming radio signal vary the amount of the corona, thus giving the necessary modulations for the picture.

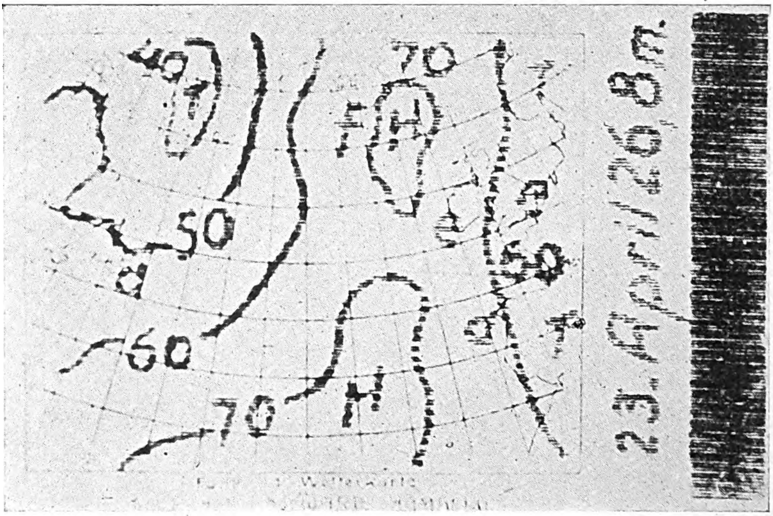

Fig. 4

Such developments are very much to the point, as surely many contributions to radio pictures will come from this stimulation; but the ease with which results will come will be with nothing like the simplicity of the broadcast reception as we know it today, where a crystal and a telephone receiver start a novice on the road to nine tube sets,—later reduced to four.

Television

Great strides have been made in the even more complicated art of television—complicated by the element of vastly greater speed. In closing this all too brief summary of the outstanding workers in these fields during the past two years, acknowledgment must again be given, as was done two years ago,[1] to the continued painstaking work of Professor A. Korn of Germany. It was he who first began to get real pictures over wire lines twenty years ago.

[1] R. H. Ranger, Proc. I. R. E., 14, 161; April, 1926.

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Photoradio Aims

As all engineers know, no particular features of their work may ever be considered as ideal solutions,—as in practical politics, it is necessary to be satisfied with compromises which conform to the general advance of the art. However, it is well to have stars on which we attempt to hang our pictures. In photoradio they have been the following:—

1. Economical operation
2. Continuous operation
3. Daylight operation
4. Visible operation
5. Finished operation

Economical operation is of course the foundation of everything worth while in the design. Naturally this economy must be in accordance with the use to which the equipment is to be put. If a larger number of sets were required, the cost of the sets themselves becomes the controlling factor as in equipment for amateurs; in photoradio to date, the prime consideration has been economy of operation—the best possible results with the least effort. Simplicity of operation is a great contributor to economy, of course, and that is where the other factors enumerated above become important. Nothing is so disastrous to production as discontinuous operation. That explains why the rotating cylinder type of equipment has been shunned in photoradio.

Daylight operation is far more satisfactory from the operator's point of view. Photography is a wonderful aid in many operations, and in much picture transmission is the obvious answer. It is certainly one of the places where compromise must be considered most carefully. The finesse of photographic recording cannot be rivalled. However, it may be said in no uncertain terms that there are many variable factors in the chemistry and physics of ordinary photography, such that its operation has become an empiric art rather than a fixed operation where definite rules will ensure a good picture in the hands of the average operator. For this reason, to add to it the furth er variables of a communication system increases the chances of failure tremendously. Therefore, certainly in all the original work over great radio distances, the greater certainty of visible and daylight operation has been depended upon. When the complete system so constituted has been operated sufficiently to reduce the uncertainties of the newer elements in the picture, it may well be that it will be advantageous to return to the use of photography with greater assurance. The

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entire matter is of course still in a state of fluctuation, working towards the best commercial method.

Furthermore, the exact requirements of photoradio on a radio circuit have brought about a tightening of the performance of that radio circuit. The ability to key a transmitter speedily and accurately is a striking example.

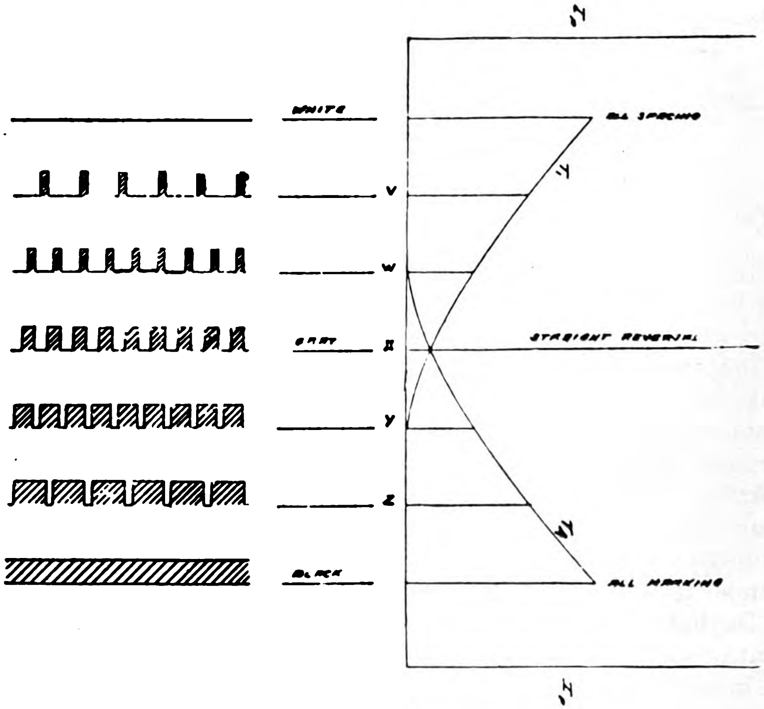

Fig. 5

It is of course advantageous that the operation be complete when the transmission has stopped. This promotes speed of delivery, which is one of the battle cries of a communication system.

Design Technique

If there is one rallying point for all of the photoradio technique, it is the use of the dot-dash method of representing picture values. This becomes at once the central feature around which all the simplification of equipment and operation is based. Variation in radio signal intensity is all too well known to require mention here. A system which is based on signal intensity to represent the picture values is placing itself under a severe handicap at the start. The dot-dash method changes light intensity variations into time variations. An isolated

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dot represents a light part of the picture, close dots represent gray, and heavy dots becoming solid dashes represent black. A scale showing this relationship is given. (Fig. 5) Such solid characters may be handled over involved links of wire lines to radio without material difficulty. It is only necessary that the radio transmitter trigger-on when a character is starting and trigger-off neatly at its termination. This simplification of operation has led to our realization of another important principle of design which we have called "inherent accuracy." It means the elimination of troublesome adjustment and compensating for large variables by making their effect nil on the result.

Another example of inherent accuracy came to light in the choice of gears in the driving mechanism. In some of the early apparatus, it was found that there was an occasional tendency to have errors in sequence in synchronizing—keeping the machines in step. The trouble was found to be in the fact that the gear ratios were such that they came out unevenly, such that the teeth would mesh in a given sequence for one line, and in an entirely different sequence for the next; and would then go back to the first sequence for the third line, and so on. In consequence, if the shaft carrying one of the gears was out slightly, the regular error would show up. The answer has proved to be to make the gear ratios even.

Push-Pull Relay

Everyone is familiar with the famous combination of "push-pull," but we have made use of the general principle in a unique way. Inertia is present in both mechanical and electrical devices. If the impulse to be conveyed is constant in frequency, such as a 60-cycle wave, the inertia factor is constant and therefore may be disregarded as far as obtaining correctly spaced impulses at the receiving end, but the whole idea of picture transmission is that there shall be change—change conveying thought in one form or another from the sender to the receiver.

The "reading condenser" has been a well known means for reducing the effect of electric inertia in d-c telegraphy. The time involved in producing a desired effect at a distance is determined by the time it takes to build up the current to a sufficient intensity. Very often, however, it is possible that the final current value may be much greater than that required to produce the start of the signaling effect. Under these conditions it is possible to dispense with some of this final current strength and use it to speed up the start and finish of the signal. A simple plan for this purpose is to include a condenser with a shunting

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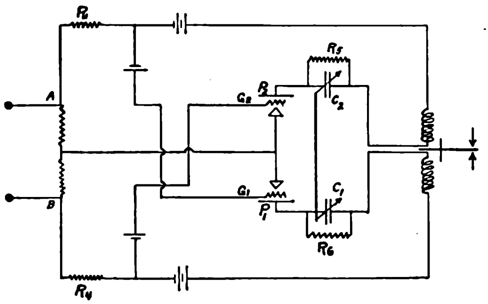
resistance in series with the main signaling circuit. This will give a large current impulse on the start and finish of the signals, and only a moderate value for the steady state.

The vacuum tube gives another simple way to reduce the inertia of inductance. By tripling the voltage on the magnetic amplifiers in connection with the long-wave stations of the RCA, and introducing twice the straight resistance of the coils, practically three times the speed of keying was obtained, with consequent improvement in the picture reproduction.

As a means of improving the quality of the impulses a broad resonance has been introduced into these inertia reducers. The particular set-up is shown in Fig. 6. Double current is used in practically all



D.C. PUSH-PULL RELAY

Fig. 6

of our d-c signaling work. Let it be assumed for the moment that a spacing current from the distant point is flowing in such a direction as to make point A positive with respect to point B. Under these conditions, grid G_1 will be positive and grid G_2 negative. Under these conditions, plate P_1 will be passing current in a direction through the relay to hold it to spacing. The plate current will carry through resistance R_6 and R_4. The current through R_4 will act to make the grid G_2 even more negative. Likewise the reduction in the plate current of P_2 due to the negative value of its grid G_2 will reduce the current through R_5 and R_1 so that the grid G_1 will be even more positive. The effect of such a hook-up is that now when the current through A-B is reduced to zero, the charge on the condensers C_1 and C_2 will be such as to cause current to flow through R_1, R_2, R_3, and R_4, which will swing the grids oppositely to their previous condition and give a marking impulse. If the current from the line should still remain at zero, the charge on the condensers would quickly dissipate itself through the resistances and the inductance of the relay coils included in the plate circuits. As soon as it did, it would swing back to the ori-

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ginal condition, and as a result a series of dots would be produced. The frequency of these dots may be set at will by proper choice of the value of the capacities C_1 and C_2 with respect to the resistances and the inductances of the relay coils.

The net effect of the action is to be ready for any change, and immediately work with it to a maximum. It accomplishes much the same effect as the famous Gulstadt relay of cable technique, but does not depend on the less reliable contacts of mechanical relays for its reversing action. The speed of the reversals of the push-pull relay is set to correspond to the fastest dots that the complete communication setup will allow. Such a push-pull relay is included both at the transmitting station and at the receiving station where the signals are finally delivered to the recording equipment.

It is hard to estimate accurately the improvement such equipment gives, due to the fact that every improvement brings out the fact that there are weaknesses elsewhere, so that it may be better said that photoradio operation throughout the past three years has been a case of gradual improvement of all the contributing factors, always hitting the weakest spots as they develop when the other factors pass them by. D. G. Ward and J. L. Finch are particularly responsible for the increased speeds that have been accomplished at the transmitting stations during this period.

Reverse Lead Screw

In order to meet the requirements of continuous operation, many plans have been proposed. Naturally the one that is most desirable is one that is continuously acting with no reversing clutches or connecting parts. On the early photoradio equipment, this was accomplished by means of a mangle rack with gear—a small ordinary gear engaging with a gear in the form of a race track. While this device has operated satisfactorily, it is not rugged and is far from inherently accurate.

In its place a reverse lead screw has been laid out. (Fig. 7) While the making of the first one was a very involved proposition, it has proved a fairly simple matter to duplicate them, as is now being regularly done in the excellent machine shop of H. O. Boehme, of New York, under the direction of Frank Kunc of our staff.

This reverse screw is cut in a solid shaft which is placed lengthwise at the base of the photoradio transmitter or receiver. A follower engages with this thread and conveys moving force to the analyzing head of the machine. The movement is uniform through 95 per cent of the travel—the remaining 5 per cent of the stroke is occupied in reversing

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the act. This does not mean that this entire 5 per cent is unserviceable for picture work, however; in fact the analyzing and recording is continuous up to the immediate end of the stroke when the overlapping interferes.

The curve for the reversing cut was first made in a very much enlarged form in steel as a master, and then by a pantograph arrangement this master was used to get the smaller cutting of the reversing section of the screw.

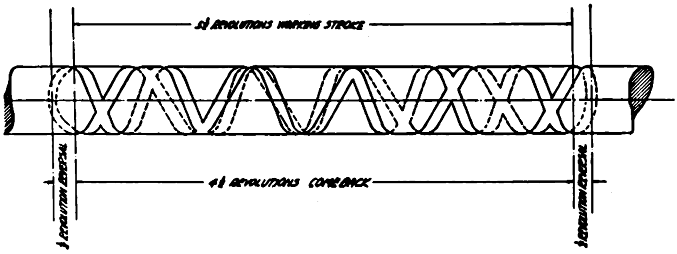

Fig. 7

Very satisfactory results from the point of service and constancy have been obtained with this screw. It is to be noted that we are still using the action of making the analyzing head work in both directions. At both transmitter and receiver, the picture making goes on while the analyzing is proceeding either from left to right or from right to left. This is inherently more difficult to do from the point of view of synchronizing, but the fact that we have done it means that we have found means of synchronizing of a fairly simple sort. If it were not for the fact that the recording is visible, however, it is certain that we would never have dared to use this back-and-forth method. As it is, we are able to correct the synchronizing very easily during the actual reception of a picture. The fact that this two-and-fro analyzing requires such rigorous synchronizing and "framing" has of course strengthened our will power in the development of synchronizing generally. Another form of speed control, which is of course the basis for synchronizing, has been developed.

Air Speed Control

Wherever radio engineers engage with problems their first solution is naturally by the use of electricity. And there is no question but that electricity is a mighty useful agent, particularly as we gain familiarity with its uses. However, with the excessive amplifications that we indulge in, we have to be very careful not to have other electric forces present which may give interference in such amplification. Speed

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control is a very good instance. In central office installation, there is not much difficulty in obtaining the necessary shielding from ordinary audio amplifications, but if it is desired to place a photoradio receiver directly next to a short-wave radio receiver for example, the problem becomes a little more involved. Of course one answer is to take the bull by the horns and provide the proper shielding and reduce the interference as much as possible. But it is well to inquire sometimes into other possibilities which inherently eliminate such interference.

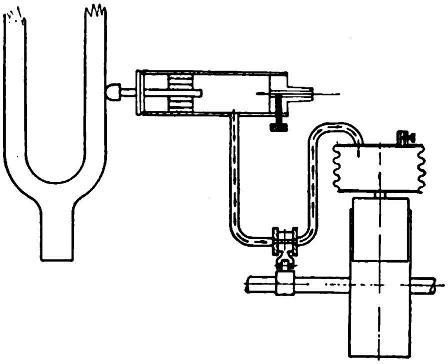

Fig. 8

As this section and the next will show, we have recently undertaken an extensive study of the uses of air. In line with this we have developed an air speed control. One form makes use of an air driven tuning fork. (Fig. 8) It is very easy to drive an air fork; all that it is necessary to do is to have an air chamber with a slightly constrained intake. This air chamber is broadly resonant to the frequency of the fork. The fork tines then have plungers on their ends which work in and out of openings in the air chamber. The action involves the filling up of the air chamber and then giving pressure to the two plungers which drive out the fork tines. In so doing they release the air pressure in the chamber quickly; the fork tines then return into the chamber, and this allows an even higher pressure to be built up in the chamber before the fork tines are again driven out to release the constrained pressure. The fork is self-starting. The frequency is modified by pressure but it is not difficult to get quite constant air pressure by the use of reduction valves, and the very fact that the frequency may be varied within narrow limits provides a means of fine adjustment. Now that we have the fork vibrating, the next question is to apply this vibration to motor speed control.

To control the motor speed, the air chamber has a small valve in it which communicates to an air brake on the motor, but the valve

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is opened by the motor once each revolution of the motor. If the valve is opened by the motor at a time when the pressure is at a maximum, a good bit of air will push through the valve to enter the brake. This will of course tend to reduce the motor speed below the point where the valve will open at a high pressure time. When it does so, the motor will again speed up and will find a position between the maximum and minimum air pressure periods where its speed will hold quite truly to the fork speed. Obviously there is nothing very electric about this to cause interference, and in fact an air motor could be used if necessary but it seems to be very easy to overcome brush trouble from motors of small size so that this has not proved necessary.

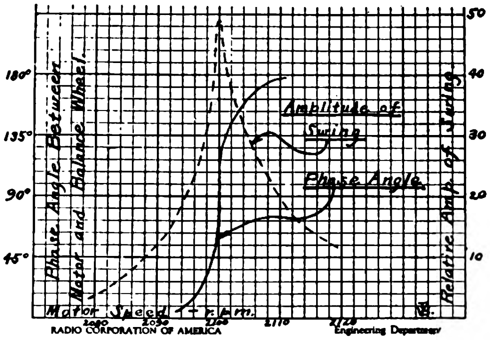

Fig. 9

Another air drive has been worked up from this by Mr. Braman and Mr. Nelson. We are all quite familiar with the use of resonance in mechanical as well as electric arrangements to effect control of one type or another. We are also quite well aware of the fact that phase differences exist, but there are not many instances where both phase and amplitude of resonance are used to effect the control. This is here accomplished. The setup consists of a cam on the motor shaft which gives the slight impulse necessary to set a vibrating arm in motion. The fact that this cam is reduced by levers to give an impulse of the order of one thousandth of an inch shows what good resonance it is possible to have in the arm. Plotting the displacement of the arm from a given reference point in the position of the motor as it revolves, Mr. Braman has developed this most interesting curve of response of the vibrator with respect to the speed of rotation of the motor. (Fig. 9) It will be appreciated that it is a combination of phase displacement and amplitude of resonance of the vibrator. It will be noted that it has a very steep portion.

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To make use of this curve, air was then driven from a nozzle through a sector carried by the vibrating arm. This air was then carried on to brake the motor. The result is a very fine setting of the speed on this frequency curve of the vibrator.

Hot Air Recording

In the search for visible, daylight, finished recording, it is believed a new departure has been realized.

We are all familiar with the sensitive gas flame of Dr. Koenig of Germany. It was this which started the development of the hot air

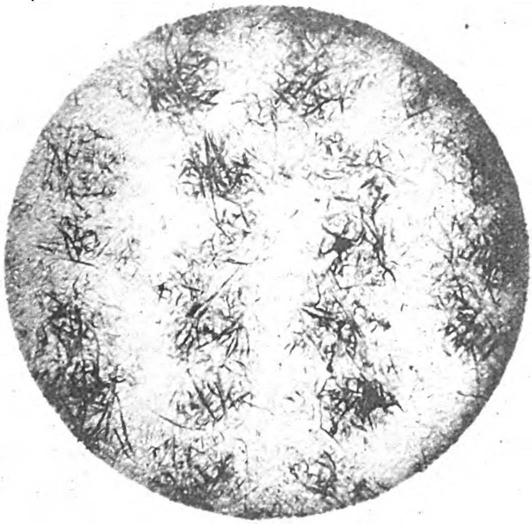

Fig. 10

recording, but as a matter of fact when the obvious attempt was made to make incoming radio signals control the sensitive flame, it was found that a much too powerful agent was at hand; it was very difficult to keep the gas from eating up everything as well as the recording paper. Then it was that Mr. Hansen worked up the use of plain hot air to do the job. Likewise, an electric spark was used, but the wear and tear on the point did not prove feasible from the operating position.

By all odds the heavy part of this new development is in the sensitive paper. We are all so familiar with extreme sensitivity of photographic paper. We have tried to duplicate this where heat rather than

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light is the sensitizing agent. Likewise we have imposed the additional restriction that the product shall be finished without further treatment.

In beginning the chemical investigation of this we had the good fortune to be able to start with R. S. Bicknell as consulting chemist. He has been seconded by Mr. Morehouse who joined our staff for this specific investigation. Mr. Morehouse gives the following general outline of his problem.

Heat-Sensitive Paper

"Photoradiograms are recorded by means of a fine jet of heated air directed against heat-sensitive paper. When the incoming radio signal calls for a dot or a dash, the hot air stream is permitted to flow against the paper; when a space is required, the hot air stream is automatically prevented from striking the sensitized paper surface. This effect is accomplished by means of an electrically operated valve, the action of which is governed by the incoming radio signal.

"The character of the record left on the paper depends on the nature of the sensitizing agent, and also on the physical character of the paper base itself. The sensitizing agent in use at the present time consists of a practically colorless mixture of chemical salts, which are capable of undergoing what is known as an endothermic double decomposition reaction, with the formation of brownish black products. In other words, until heat is applied to these salts they remain in contact with each other on the surface of the paper in a comparatively inert condition, retaining their colorless appearance. But when heat is applied they absorb it rapidly, become mutually interactive, and form the black products which we see as a dot or dash on the picture.

"The importance of the physical character of the paper stock is best brought out with the aid of photomicrographs. Fig. 10 shows a few dots made by the hot air stream on an ordinary uncoated book paper made sensitive to heat. It will be noticed that the dots have a very pronounced fibrous or stringy character. The same fibrous appearance is noticeable in Fig. 11 which shows dashes made on this same type of paper. Fig. 12, on the other hand, shows the type of dot obtained with a clay-coated paper similarly sensitized. In this case the dots have a much softer and smoother appearance because of the fact that the paper fibres are completely covered by the thin layer of clay. Fig. 13 shows dashes made on the same clay-coated stock. In all four cases the chemical composition of the sensitizing agent is the same; the difference between the first two and the last two is entirely due to the difference in the physical character of the paper surface. In cases three and

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four the effect is very similar to that obtained by using photographic paper in the process of recording with the aid of light instead of heat."

OPERATION

We now come to the question of operating the equipment. Commercial picture transmission was inaugurated on May 1st, 1926, between London and New York, and likewise it was established on the West Coast between San Francisco and Honolulu. There was considerable excitement incident to the inauguration of the service, especially from London to New York. It was finally determined that one picture only to each newspaper and commercial concern would be sent in the sequence that they were received for the inception of the service. After the novelty had worn off, the photoradio service came into extensive use on a business basis.

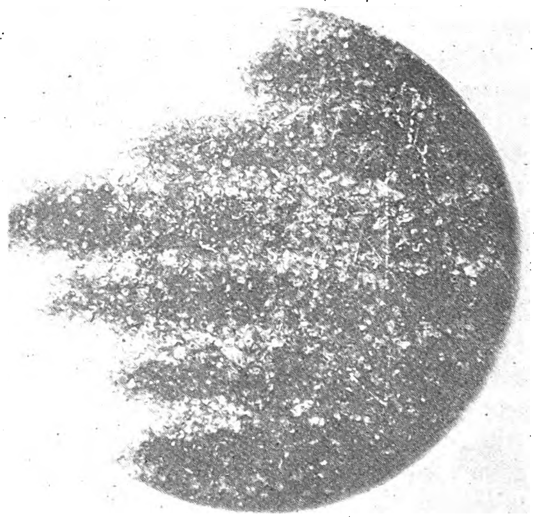

Fig. 13

Besides the news service, however, there has been a gradual but general extension of the service first into the style and fashion field and then into banking and commercial operations.

A few months ago for example, a bank came to us with the request to transmit the three signatures necessary to authorize a certain commercial transaction involving one million dollars. They advised that

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there were but three days remaining to get these signatures into London, and as there was no Lindbergh expedition scheduled for that particular day, it would be impossible to get these signatures across by any other means. However, we were told to take our time in the three days, and get a good facsimile of the signatures across.

We advised London of the situation and asked them to work with us until a satisfactory facsimile had been transmitted. The equipment at each end was set in motion; in twenty-five minutes (it was a half-size picture) transmission was put through once. London was asked for a report on this first transmission with the idea that the reply would indicate that it should be made darker or lighter or something else on the second trial. Back came the laconic reply "O.K.," and that was all that was necessary of the three days' grace to complete the million dollar transaction, as far as photarodio was concerned.

Department Store Demonstrations

At the request of Kauffman's of Pittsburgh we embarked on what might be termed barn storming expeditions. It might seem at first that such excursions would be far afield from a technical development, but there is little question but that these demonstrations have speeded up the development of photoradio to commercial reality in great degree. The cost of these demonstrations was met by the department stores. It therefore became imperative that a very worth while and useful service be given which would be capitalized by the department store.

The basis of these demonstrations was to bring the latest styles from Paris and London directly into the department store. The transmission to New York from Europe was easy—the problem came in relaying this on to the stores. Short waves were used for the relaying. WIZ at New Brunswick and WAQ of the Westinghouse Company at Newark were used for this relaying service; they were connected with Broad Street, New York by land lines; the picture reception comes in over land lines from Riverhead, Long Island and Belfast, Maine, to Broad Street, which therefore may be termed photoradio central. The well known fading propensity of short waves became of less importance to the photoradio signals on the dot-dash plan. Added to this, limiting was used with great success. Further demonstrations were then held at Strawbridge and Clothiers in Philadelphia, Jordan Marsh in Boston, Marshall Field in Chicago, and lately at L. S. Ayres. Further barn storming could have been developed, but it was felt that with successful termination of these demonstrations their purpose and usefulness as far as advancing the art is concerned was sufficient.

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Navy Development

As the request of the Navy Department, successful transmissions were accomplished of navy manoeuvers in Honolulu direct to New York. Subsequent to this, the Navy Department has installed photoradio equipment at the Bureau at Washington. Transmission and reception has been successfuly accomplished between the photoradio station and the Radio Corporation stations. Likewise, condensed equipment has been installed on board the USS *Seattle*, later transferred to the *Texas*, when the latter became the flagship. Perhaps the most successful transmission with this equipment has been an entire page of printed matter from New York to the *Texas* when she was lying off San Diego something like three thousand miles from New York.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

GRID LOSSES IN POWER AMPLIFIERS*

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Summary—*In spite of the widespread use of radio-frequency power amplifiers, apparently no measurements of the power required to drive such amplifiers have ever been made. In this paper, a study of driving power is made at 60 cycles. It is found that the power input to the grid is proportional to the d-c grid current raised to the 1.34 power. The proportionality factor depends on the type of tube and the plate circuit conditions. With constant grid current the grid power input is found to be practically independent of grid bias voltage. A simple theory, based on an assumed relation between grid current and voltage, agrees quite well with the experimental results. Data are then given on six types of commercial air cooled transmitting tubes showing driving power, a-c driving voltage, gross power output and efficiency as functions of the d-c grid current. Finally, the effects of primary and secondary electron emission by the grid on the driving power are considered.*

1. Introduction

HIGH-VACUUM electron tubes of the transmitting type have taken on a steadily increasing significance ever since their introduction into the art of radio communication about a decade and a half ago. During this period, numerous technical investigations into the operation of such tubes as radio-frequency power amplifiers and oscillators have been published. These have dealt largely with the conditions under which optimum conversion of plate supply power to radio-frequency power is obtained. Thus the plate circuit has received careful consideration. The grid circuit, on the other hand, has been almost completely neglected. The object of this paper is to present measurements of grid circuit losses on a series of typical air cooled transmitting tubes. A knowledge of such losses is particularly useful for the design of multistage amplifiers, where the plate circuit of one tube must supply the grid circuit losses of the succeeding tube.

2. Operation of Radio-Frequency Power Amplifiers

The elementary radio-frequency power amplifier circuit is shown in Fig. 1. A voltage E_g, which generally approaches a sine wave (and is so assumed throughout this paper), is impressed upon the amplifier in series with a bias voltage E_c. Since maximum power output is desired, no particular attention is paid to the wave shape of plate

* Dewey decimal classification: R342.5. Original manuscript received by the Institute, March 21, 1929.

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current, a tank circuit L,C,R being relied upon to filter harmonics out of the output. This circuit is tuned to the frequency of E_g and presents a load impedance of very nearly L/RC ohms at unity power factor to the fundamental component of plate current. The impedance of this circuit for harmonics is usually quite small in comparison with L/RC, hence, in spite of badly distorted plate current wave shapes, the voltage across the tank circuit is practically sinusoidal and 180 deg. out of phase with E_g. The lower the decrement (or ratio of resistance to reactance) of the tank circuit, the more nearly a sinusoidal wave shape is obtained.

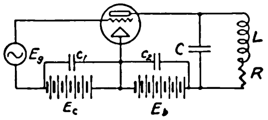

Fig. 1—Elementary Radio-Frequency Power Amplifier Circuit

Now it is well known that for high plate circuit efficiency, plate current should flow during less than half a cycle of the exciting voltage.[1,2,3,4] This is accomplished by making E_c sufficiently negative so that if E_g were reduced to zero, no plate current would flow. For proper excitation, E_g must now have an amplitude sufficient to drive the grid positive, at the instant of maximum grid potential, by an amount which varies between 50 per cent and 100 per cent of the *instantaneous* plate potential at that moment. Unless this condition obtains, an inefficient transfer of plate current results. This maximum positive grid potential may lie between 5 per cent and 15 per cent of the plate supply voltage. During the time that the grid is positive with respect to the filament, an electron flow to the grid takes place. This flow causes a twofold loss of power; first, a loss due to electron bombardment of the grid and second, a charging of the bias battery E_c. These are the losses which form the subject of this investigation. Any losses caused by the grid current in transformers, condensers, or other apparatus associated with the grid circuit will not be considered here because they are primarily a matter of circuit design.

3. Apparatus

Measurements at radio frequencies are rather difficult, and one of the most useful of all electrical instruments, the commercial low-frequency oscillograph, cannot be applied. Except for extremely high

[1] H. Rukop, *Jahr. d. Draht. Tel. u. Tel.*, 14, 110, 1919.
[2] J. H. Morecroft, *Jour.* A. I. E. E., 38, October, 1919.
[3] Marius Latour and H. Chireix, Proc. I. R. E., 11, 551; October, 1923.
[4] H. G. Moller, "Die Elektronenröhren," Vieweg and Sohn, 1922.

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frequencies, i.e. of the order of 3×10^8 cycles, electron tubes show no time lag; consequently, for the present purpose, the same results can be obtained at commercial power frequencies as at radio frequencies. For these reasons, all experiments reported here were made at 60 cycles.

The circuit arrangement employed is shown in Fig. 2. Instead of tuning the tank circuit, L-C_2, which would have been rather difficult at the low frequency and high plate voltages used, the tank was constructed of a fixed inductance and capacitance and the frequency of the sine wave alternator varied until the plate current was a mini-

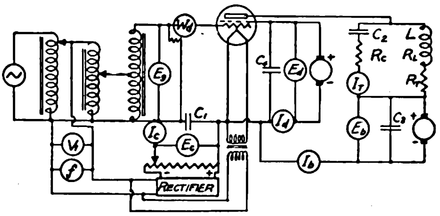

Fig. 2—Test Circuit

mum. Experiments showed that when the plate current is a minimum, the plate efficiency is maximum. The voltage at V_1 was maintained at 110 volts for supplying the frequency meter and grid bias rectifier. The grid exciting voltage was supplied by a 5-kw transformer so that the asymmetrical grid current would not distort the actual grid voltage. Oscillograms showed that the instantaneous grid voltage was for all practical purposes the sum of a pure sine wave and a constant voltage.

The tank circuit consisted of an air core inductance (L) of 0.674 $\pm$0.001 henries and 5.30 ohms d-c resistance, and a bank of high voltage oil condensers (C_2) whose capacity and series resistance at 60 cycles were 10.23 μf and 1.95 ohms, respectively. The constants of the condensers were determined by taking a series resonance curve. In this manner the frequency for equal inductive and capacitative reactances was found to be 60.56 cycles per second.

For varying the load on the tube, additional resistance R_t could be connected into the tank circuit. If the frequency is adjusted until $\omega^2 LC_2 = 1$, the effective resistance of the tank circuit is

$$\frac{R_c(R_1+R_t)+\dfrac{L}{C}}{R_c+R_1+R_t}$$

Since in the case under consideration $R_c = 1.95$ ohms, $R_l = 5.3$ ohms, $L/C = 65{,}900$ (ohms)^2, the first term of this expression is negligible ex-

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cept for values of R_t of the order of hundreds of ohms. The largest value of R_t used here was 61 ohms. Hence, the effective resistance of the tank circuit is simply L/RC ohms, where R is the total resistance of the tank circuit, or $R_l = R_l + R_c + R_t$.

The gross output of the tube is $I_t^2 R$. For the larger values of R_t this expression gives output figures which are higher than the true output, because harmonics of the plate current flow largely through the capacity branch of the tank circuit. This was avoided by switching the I_t meter from the capacity to the inductance branch of the tank circuit (see Fig. 2). Corrections for the plate current were then necessary.

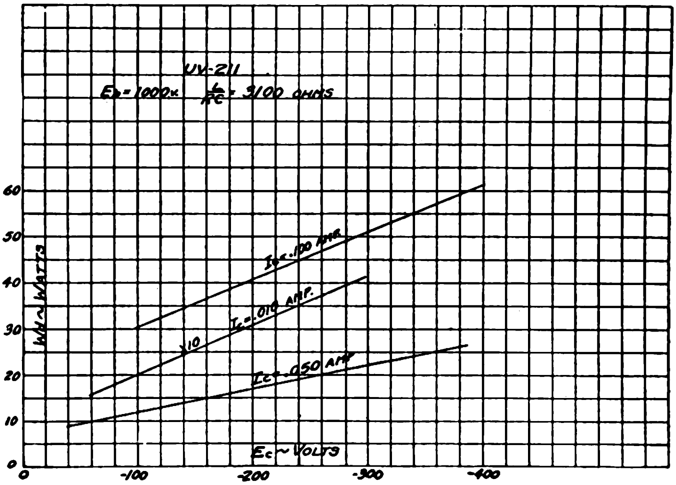

Fig. 3

The filament was supplied from the same source as the grid excitation voltage. The effect of this was to increase or decrease the grid excitation by an amount equal to about half the filament voltage. In all cases the filament voltage was adjusted to the rated value. On account of the excess emission available in thoriated filament transmitting tubes, variations in filament voltage up to ±10 per cent have practically no effect on the output. This was convenient as it made accurate adjustment of filament voltage unnecessary.

Grid driving power was measured by a wattmeter connected to the grid circuit as shown in Fig. 2. A special meter with a current coil resistance of 0.9 ohm was selected. Low current coil resistance is essential if distortion of the grid voltage wave form is to be avoided.

The arrangement for supplying the voltage E_d in Fig. 2 was only used when four-element tubes were being tested.

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4. Experimental Results

The power required to drive the grid of a power amplifier can be divided into two distinct components. With sinusoidal grid excitation as soon as $\sqrt{2}\, E_g > E_c$, grid current flows during the portion of each cycle when the grid is positive with respect to the filament. This current must overcome two voltages—the voltage from grid to filament and the bias voltage E_c. The average power loss in the bias source is simply $E_c \cdot I_c$, assuming that C_1 by passes the radio-frequency components of grid current and has negligible series resistance. If we call the average driving power W_d and the average grid loss W_g, we get

$$W_d = W_g + E_c I_c. \tag{1}$$

Now W_g is a function of E_g and E_c, but since I_c is a function of these same variables, we can eliminate E_g and consider $W_g = f(E_c, I_c)$.

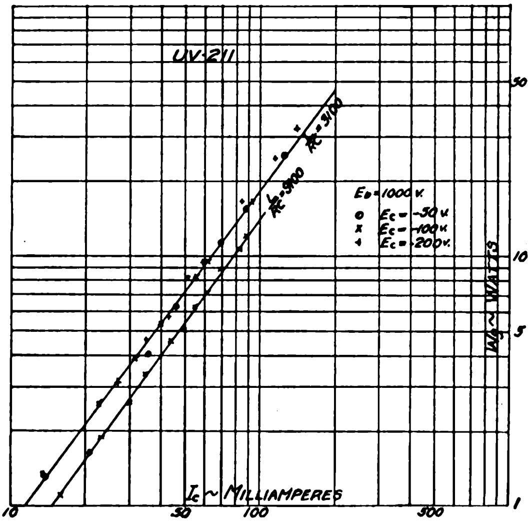

Fig. 4

The first measurement of W_d was made with a UV211 operated under "full load" plate circuit conditions. The method of determining these conditions will be described later. E_g and E_c were varied in such a manner as to keep I_c constant. The result is shown in Fig. 3. These curves show that for $I_c = 0.100$ and $I_c = 0.050$

$$\frac{\partial W_d}{\partial E_c} = I_c.$$

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For $I_c = 0.010$ ampere $\partial W_d/\partial E_c = 0.91\ I_c$. However, this last value of I_c is considerably below the normal operating value for the UV211. In the normal operating region we can say $\partial W_d/\partial E_c = I_c$. From (1) it is evident that this can only be true provided $\partial W_g/\partial E_c = 0$; that is, the grid loss is independent of the grid bias for constant I_c. Equation (1) thus becomes

$$W_d = f(I_c) + E_c I_c. \tag{2}$$

This independence between W_g and E_c was also observed in the case of other tubes. It is only necessary, therefore, to show W_g as a function of I_c.

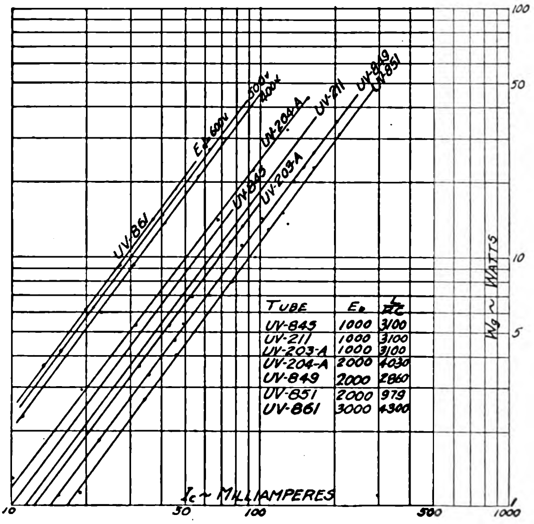

Fig. 5

The upper curve on Fig. 4 shows $W_g = f(I_c)$ for a UV211, taken with several values of E_c. The points are all on or close to one straight line. Deviations were found by repeated measurements to be due mostly to experimental error resulting from variations in supply frequency and voltages. From this curve we get

$$W_g = 0.039 I_c^{1.34} \text{ watts} \tag{3}$$

where I_c is expressed in milliamperes.

So far, full load plate circuit conditions were maintained. These are the most common conditions. The effect of a change in the plate load can be seen from the lower curve of Fig. 4 which shows W_g for a light load curve. The coefficient in (3) is now reduced to 0.03, which is a reduction of 23 per cent. It is easy to see why the grid loss should

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be less for light loads than for heavy loads with a given value of I_c. As the plate load resistance L/RC is increased, the minimum instantaneous plate voltage decreases steadily and may even go negative. This means that the maximum positive grid voltage may be considerably above the plate voltage. Under these conditions the grid collects most of the electrons emitted from the filament. Thus other conditions remaining the same, as the plate load resistance is increased I_c will increase. E_g must now be decreased to obtain the same value of

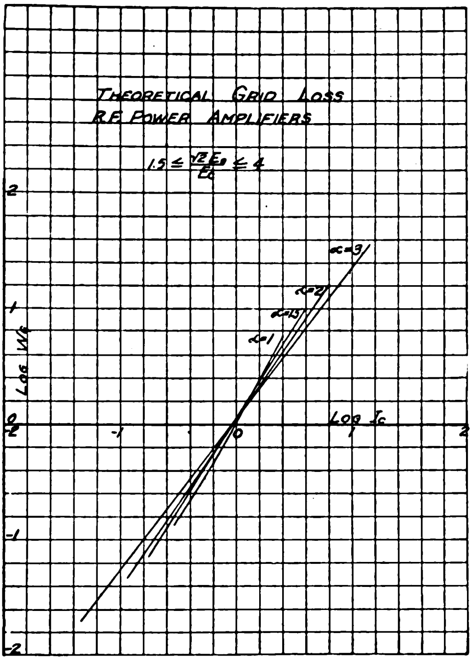

Fig. 6

I_c as at full load. After this change has been made, the wave shape of grid current is not much different from what it was originally. Consequently, since the decrease in E_g decreased the positive grid voltage, the grid loss is less.

$W_g = f(I_c)$ was now taken for a series of seven different types of transmitting tubes[5] under full load conditions. These results are shown on Fig. 5. For every tube the function $W_g = f(I_c)$ is expressible as

$$W_g = AI_c^{1.34} \text{ watts} \tag{4}$$

where I_c is again in milliamperes. The value of A is given in Table 1.

[5] Technical information on these tubes is given in Appendix III.

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TABLE 1

Tube	*L/RC* ohms	*A*
UV845	3100	0.0467
UV211	3100	0.0390
UV203A	3100	0.0356
UV204A	4030	0.0531
UV849	2860	0.0293
UV851	979	0.0244
UV861 ($E_d = 500$ v)	4300	0.107

No quantitative relation connecting the coefficients A and the dimensions of the tubes could be found. However, a little consideration will show that the following qualitative statements must be true. Other conditions remaining the same, as the grid mesh is increased, (thus increasing the μ of the tube) A must decrease. This agrees with the experimental results in the case of the UV845, the UV211 and the UV203A which differ only in the grid mesh and have values of μ of 5, 12, and 25, respectively. As the μ is increased by moving the grid toward the cathode, A must decrease.

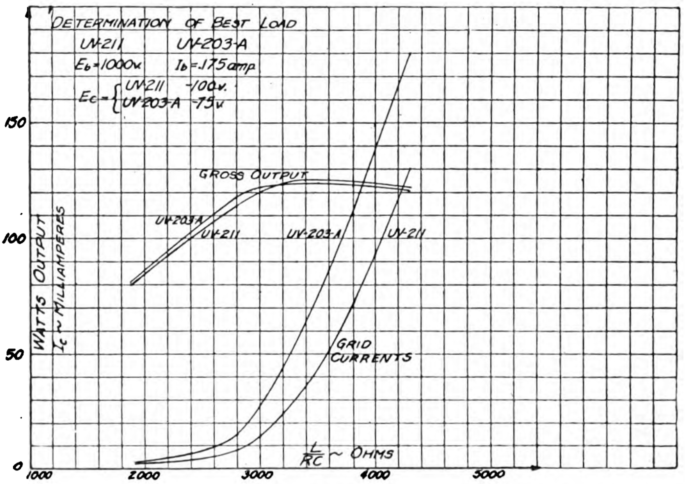

Fig. 7

5. Theory of Grid Loss

The maximum error in the exponent 1.34 was estimated to be about ±0.04. Thus the extreme values of the exponent are 1.30 and 1.38. It is now of interest to see if this exponent can be justified theoretically. To do this it is necessary to know the function connecting instantaneous grid voltage and grid current. Analysis of oscillograms showed that

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this function is quite complicated under normal power amplifier conditions. Among other things, it is affected by secondary electron emission, which is generally present. Assuming

$$i_g = Be_g^{\alpha} \tag{5}$$

the exponent α varied from a value 1 at the lowest grid currents to a value 4 at the highest currents.

From (5) expressions for I_c and W_g can be obtained. These are

$$I_c = \frac{B}{\pi} E_c^{\alpha} \int_0^{\cos^{-1}(E_c/\sqrt{2}E_g)} \left(\frac{\sqrt{2}E_g}{E_c} \cos x - 1 \right)^{\alpha} dx \tag{6}$$

$$W_g = \frac{B}{\pi} E_c^{\alpha+1} \int_0^{\cos^{-1}(E_c/\sqrt{2}E_g)} \left(\frac{\sqrt{2}E_g}{E_c} \cos x - 1 \right)^{\alpha+1} dx. \tag{7}$$

If E_g is eliminated from these equations, W_g can be expressed as a function of E_c and I_c. Taking $E_c = 1$, $B = \pi$, $\log W_g = f(\log I_c)$ was calculated for $\alpha = 1$, 1.5, 2, and 3 for the practical range $1.5 \leqq \sqrt{2}E_g/E_c \leqq 4$.

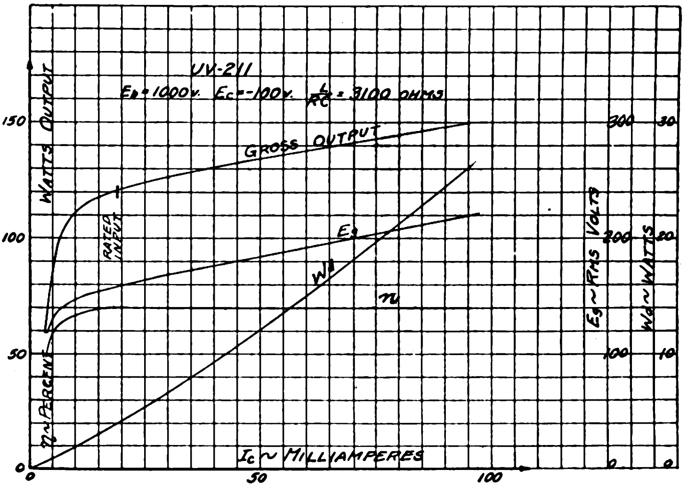

Fig. 8

These results are shown in Fig. 6. For $\alpha = 1$ a slightly curved line is obtained, but for the remaining values of α the line is straight for all practical purposes. These curves are exactly like the ones found experimentally. The slope of the experimental curves is 1.34 ± 0.04. The slope of the theoretical curves is seen from the following:

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α	Slope $=\dfrac{\partial(\log W_g)}{\partial(\log I_c)}$
1	1.73
1.5	1.57
2	1.46
3	1.30
Experiment	1.34 ±0.04

Thus even though the simple relation (5) does not hold exactly, the exponent α can vary over a considerable range and still result in a type of grid power characteristic very nearly the same as the experimentally observed characteristics.

It now remains to explain the observed fact that W_g is practically independent of E_c as long as I_c is constant. The effects of variations in E_c can be seen from (6) and (7) if E_g/E_c is held constant.

$$\log I_c = \alpha \log E_c + K_1$$

$$\log W_g = (\alpha+1) \log E_c + K_2$$

Now if E_c is increased from E_c to $m\,E_c$, the corresponding increments in log I_c and log W_g are

$$\Delta \log I_c = \alpha \log m$$

$$\Delta \log W_g = (\alpha+1) \log m$$

From these equations it is clear that if

$$\text{Slope of } [\log W_g = f(\log I_c)] = \frac{\partial(\log W_g)}{\partial(\log I_c)} = \frac{\alpha+1}{\alpha} \tag{8}$$

any increase in E_c will merely move a point up along the characteristic. Thus for all values of E_c only one characteristic will be obtained. The degree with which condition (8) is satisfied by the theoretical curves of Fig. 6 can be seen from the following:

α	$\dfrac{\partial(\log W_g)}{\partial(\log I_c)}$	$\dfrac{\alpha+1}{\alpha}$
1	1.73	2
1.5	1.57	1.67
2	1.46	1.50
3	1.30	1.33

Condition (8) is thus nearly satisfied for $\alpha=2$ and $\alpha=3$. Hence, for both these values of α, W_d is practically independent of E_c, for constant I_c. Again, there is a satisfactory agreement between theory and experiment.

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6. Relation between Driving Power and Plate Circuit Output

The measurements of grid losses given in the preceding section are most useful when accompanied by plate circuit output and efficiency. The purpose of this section is to give these additional data and to show how advantageous operating conditions can be chosen.

For economic reasons, transmitting tubes are usually operated under maximum rated conditions. All the tubes used here have d-c plate current and plate voltage limits for r-f power amplifier or oscil-

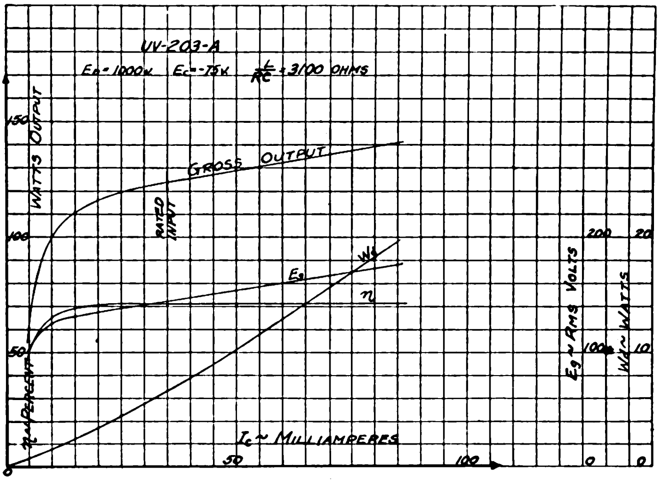

Fig. 9

lator services.[6] Thus the maximum plate input is fixed in each case. If a tube is operated at its maximum plate voltage, the input can be controlled by means of the plate load resistance L/RC, the excitation current I_c, and the grid bias E_c. The problem of finding advantageous operating conditions then consists of determining values of these three variables which will give low driving power and high plate efficiency.

It is well known that for a fixed plate current, the more negative E_c the greater the efficiency of the plate circuit. In the practical range of values of E_c the rate of increase of efficiency with E_c is quite low. Experiment shows that as E_c is made more negative, I_c must also be increased to hold constant plate current. From the preceding sections it is obvious that the driving power will therefore increase more than proportionally to E_c. The best value of E_c will therefore depend largely

[6] See Appendix III.

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on the driving power available. It was sufficient for the present purpose to choose values of E_c which gave within a few per cent of the maximum full load efficiency that could be obtained with very large values of E_c.

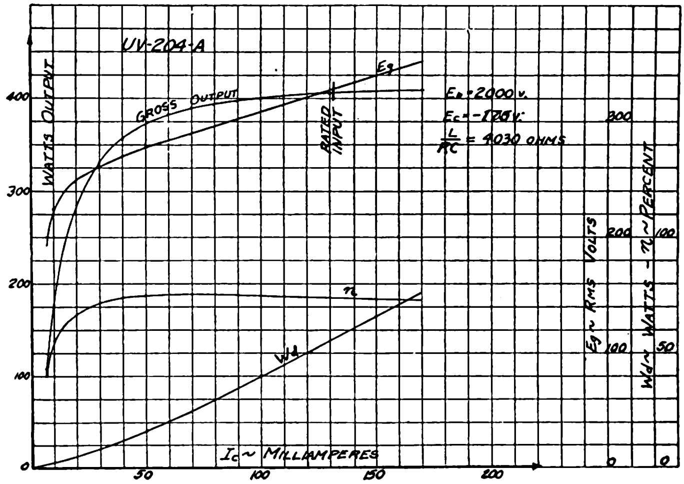

Fig. 10

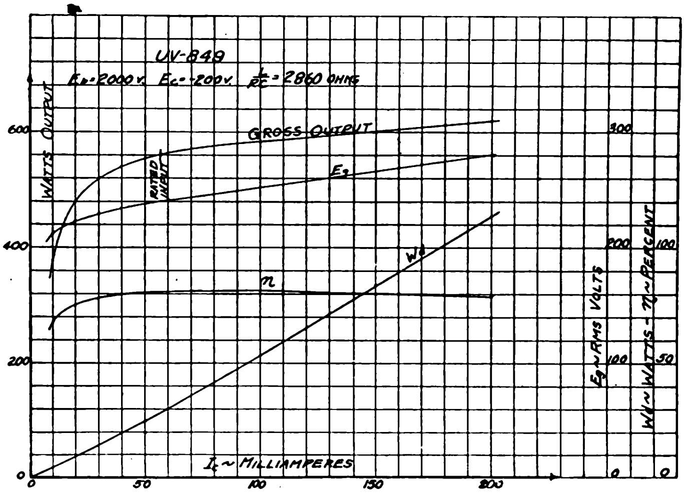

Fig. 11

Now with E_c fixed, only one value of I_c will give the desired plate input for any given value of L/RC. In this manner curves of power output against L/RC can be obtained. Two typical curves for the UV211 and UV203A are shown on Fig. 7. These curves show that

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maximum output can be obtained at about $L/RC=3100$ ohms. For higher values of L/RC the grid currents, and hence the driving power, increase very rapidly. The reason for the increased grid currents is as follows: When L/RC is large the minimum plate voltage is small. This produces a deep saddle in the plate current wave form, which loss in current must be made up by spreading the plate current over a large portion of each cycle (since we are holding a constant I_b). This can only be done by a large increase in E_g, so that I_c increases rapidly too. Thus for low driving power, L/RC should be chosen as low as possible. At the lower values of L/RC the output decreases rapidly because in this case the minimum plate voltage is high, which results in high plate loss and therefore a lower output. So when the optimum of output is broad, it is best to choose the lower values of L/RC.

The operating conditions chosen for several types of tubes are shown in Table II.

TABLE II

Tube type	E_b v	I_b a	E_c v	I_c ma	E_g v	L/RC ohms	Gross Output watts	η Per cent	W_d watts	$E_b/2I_b$ ohms
UV-211	1000	0.175	−100	19	160	3100	122	69.7	3.9	2860
UV-203A	1000	0.175	−75	35	145	3100	123	70.3	6.7	2860
UV-204A	2000	0.275	−175	131	328	4030	405	73.6	56.	3640
UV-849	2000	0.350	−200	57	240	2860	562	80.2	29.	3500
UV-851	2000	0.900	−200	300	274	979	1420	78.5	111.	1110
UV-861*	3000	0.350	−200	40	455	4300	665	63.5	22.5	4290

* $E_d=500$ v

The approximate value of L/RC for any case can be obtained in the following manner:

Let $K=E_t/E_b$. Then

$$\eta E_b I_b=\frac{(KE_b)^2}{\dfrac{L}{RC}}$$

or

$$\frac{L}{RC}=\left(\frac{K^2}{\eta}\right)\frac{E_b}{I_b}.$$

Assuming $K=0.60$, $\eta=0.70$ as average values

$$\frac{L}{RC}=\frac{E_b}{2I_b}. \qquad (9)$$

This approximation is especially convenient when tubes are to be operated at unusual plate voltage and current. Values of $E_b/2I_b$ are tabulated in Table II for comparison with L/RC.

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The first two tubes in Table II differ only in grid mesh, the UV211 having an amplification factor of 12 and the UV203A, 25. For the chosen conditions the UV211 requires somewhat less grid power. No absolute conclusions can be drawn from these data, however, as a slight change in L/RC can easily make the UV211 take more driving power.

The same things are true to a certain extent of the other tubes listed in Table II. The UV204A, for example, requires 56 watts driving power. If the L/RC is lowered from 4030 to 3810 ohms, the driving power is reduced to about 30 watts, with an attendant decrease in plate efficiency. In a similar manner, the driving power for the other

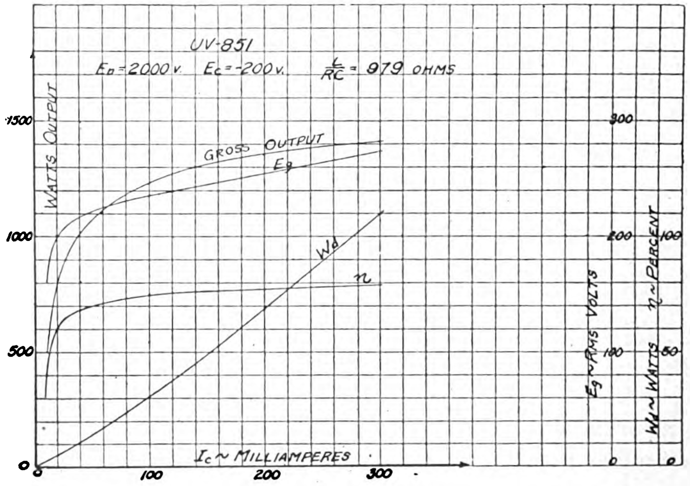

Fig. 12

tubes can be decreased. This brings out the fact which has already been explained and must be emphasized again, namely, that the operatating conditions of Table II are not the optimum conditions for all cases. These tubes can be operated very satisfactorily under conditions differing considerably from those set down in Table II. The latter are simply representative conditions which have been chosen in an endeavor to keep the driving power as low as possible without too great a loss in plate efficiency. If in any particular installation more driving power is available, it is advantageous to increase L/RC and E_c thereby increasing the plate efficiency.

Using the values of L/RC and E_c given in TableII curves of power output, efficiency, driving power and a-c grid voltage were taken as functions of I_c. These are shown in Figs. 8 to 14 inclusive. It might be argued that these curves would be more properly shown as functions

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of E_g, since E_g is a true independent variable. E_g is difficult to measure at high frequency and the provision for its measurement in actual installations, such as radio transmitters, is unheard of. I_c, on the other hand, is easy to measure and meters for its measurement are generally provided in practice.

There is nothing new or unusual about the curves of Figs. 8 to 14 with the exception that they show driving power. The reasons for the various forms of the curves are so obvious that they will not be recited here. These curves show clearly what happens when the excitation current (I_c) is varied, other conditions (L/RC, E_b, E_c) remaining con-

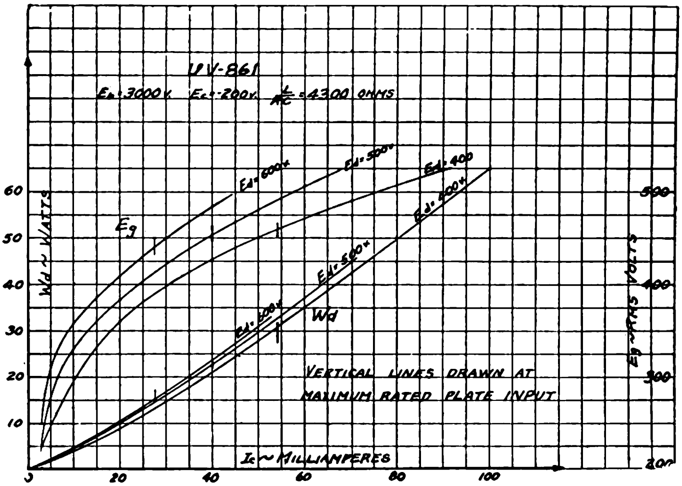

Fig. 13

stant. The output curves have a shape similar to magnetic saturation curves for iron. For this reason the value of I_c which gives an output just above the knee of the curve is sometimes called the "saturation grid current." This is generally a rather indefinite value. Furthermore, it is different for every plate load (L/RC). For large values of L/RC the saturation grid current is low and vice versa. Thus, unless L/RC is specified, the term "saturation grid current" has little significance.

Figs. 10 and 11 give an interesting comparison between the UV204A and the UV849. The latter tube is of later design and is definitely superior to the UV204A as far as excitation voltage and efficiency are concerned, when both are operated at their maximum rated plate inputs. If now the input to the UV849 were reduced to the same value as that of the UV204A by increasing L/RC its efficiency would increase

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while the driving power and a-c grid voltage required would decrease, thus still further increasing the superiority of the UV849.

Figs. 13 and 14 are some curves on the UV861 screen-grid tube as an r-f power amplifier. These were taken with the screen maintained at filament a-c potential by a 115-μf condenser. From these curves a comparison can be made when the tube is operating under various screen voltages, E_d. These are given in Table III.

TABLE III

$E_b = 3000$ v		$E_c = -200$ v		$I_b = 0.350$ a	
E_d	I_c	E_g	W_d	Output	η
v	ma	v	watts	watts	Per cent
400	54	460	30.5	678	64.6
500	40	453	22.5	665	63.5
600	27.5	440	15.0	632	60.5

Oscillograms[7] showed that under each of these conditions the minimum plate voltage was very close to the screen potential. This caused a

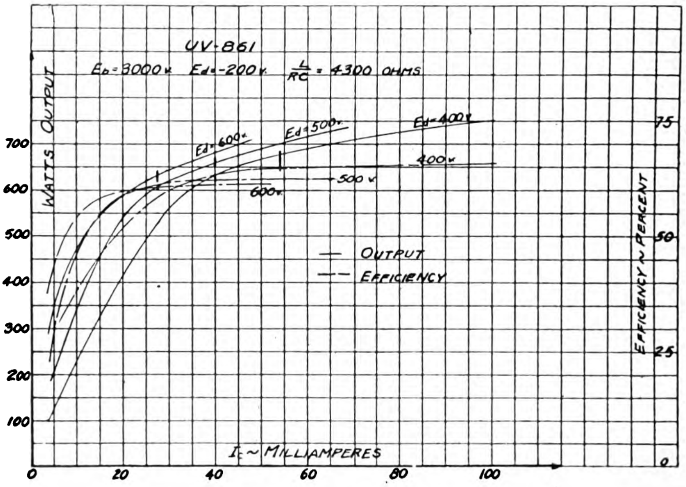

Fig. 14

deep valley in the plate current wave form. As the screen potential is lowered this valley is made less deep. Hence, for constant plate current the wave can be narrowed, thus increasing the plate efficiency. At the same time the grid must be driven more positive to get the required plate current through the screen grid. Thus for decreasing E_d, W_d increases.

7. Effect of Primary and Secondary Electron Emission from the Grid

The fact that conditions are favorable for the emission of secondary electrons by the grid, during the portion of each cycle when the

[7] See Appendix II.

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grid is positive, has already been mentioned. Suppose a certain tube does not show secondary emission in the operating range of grid voltage. Let the grid current be represented by wave *c* in Fig. 15(a). W_d is the product of the fundamental component of this wave which is in phase with E_g, by E_g. If, as in Fig. 15(a), $e_g=\sqrt{2}E_g \sin \omega t - E_c$ then,

$$W_d = \frac{E_g \omega}{\sqrt{2}\pi}\int_0^{2\pi/\omega} i_g(t) \sin \omega t \cdot dt. \tag{10}$$

If, in some manner, the grid can be activated to emit secondary electrons, so that the wave *c* is changed to *d*, the value of the integral in (10) will obviously be diminished, thus reducing W_d, the driving power. At the same time, the plate current and power output will be increased to some extent. Therefore, the introduction of secondary emission will reduce the driving power without reduction of plate circuit output. Ample confirmation of this was found in several cases. One tube in particular, when not excited to full output, actually showed negative driving power and grid current. In other words, the grid was delivering rather than absorbing power.

A limited amount of secondary grid emission is desirable in r-f power amplifiers provided it can be controlled, but when the emission reaches such magnitude as to produce a reversal of grid current as the grid voltage passes a certain positive value, other disturbing factors appear. If a grid leak is used for bias, the grid may suddenly

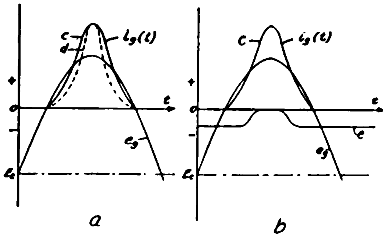

Fig. 15

assume a high positive potential which produces a destructive flow of plate current. This has for many years been termed "blocking."

Sometimes a small amount of primary (thermionic) emission from the grid is encountered. Since the grid temperature can hardly follow radio-frequency variations of E_g, the primary emission current is a continuous one provided the grid is always sufficiently negative with respect to the other electrodes to allow this current to be drawn away from the grid. From (10) it is evident that the addition of a continuous

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current to i_g (t) has no effect on the value of the integral and hence on W_g. If the grid potential is more positive than the plate during a fraction of a cycle, the primary emission current cannot leave the grid during this period. This is shown in Fig. 15(b), where the normal grid current and primary emission current are drawn separately as

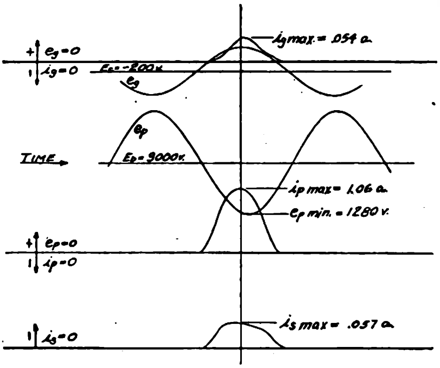

Fig. 16—UV861 under-excited.

$E_b = 3000$ v	$E_c = -200$ v
$I_b = 0.240$ a	$I_c = 10.4$ ma
Input $= 720$ watts	$E_g = 335$ v
Output $= 352$ watts	$E_d = 500$ v
Efficiency $= 49$ per cent	$I_d = -18.0$ ma

curves *c* and *e* respectively. Both have the same general phase. Hence, if variations in the primary emission current are taken into account, W_d will be increased by (10). Therefore, if primary grid emission is introduced into a tube, all other conditions remaining the same, either the driving power is not affected at all, or it is increased.

Appendix I

Symbols

E_b—d-c plate voltage
e_p—instantaneous plate voltage
I_b—d-c plate current
i_p—instantaneous plate current
E_t—r.m.s. voltage across tank circuit
E_g—r.m.s. grid driving voltage
e_g—instantaneous grid voltage
E_c—d-c grid bias voltage
I_c—d-c grid current or "excitation" current
i_g—instantaneous grid current
I_d—d-c screen current

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i_s—instantaneous screen current
$\omega = 2\pi f$—angular velocity
f—frequency, cycles per second
L—inductance in tank curcuit
C—capacity in tank circuit
R—total resistance in tank circuit
η—plate efficiency
W_d—average grid driving power in watts
W_g—average grid loss in watts

Appendix II

Since screen-grid transmitting tubes have been developed rather recently,[8] a few oscillograms illustrating their action may not be out of place here. These oscillograms were taken in the circuit of Fig. 2, using a UV861.

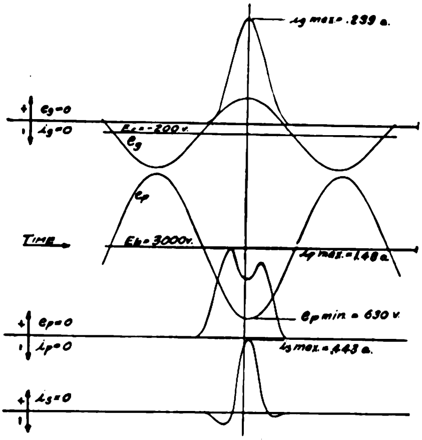

Fig. 17—UV861 fully excited.

$E_b = 3000$ v	$E_c = -200$ v
$I_b = 0.358$ a	$I_c = 48.0$ ma
Input = 1074 watts	$E_g = 496$ v
Output = 685 watts	$E_d = 500$ v
Efficiency = 63.7 per cent	$I_d = +49.0$ ma

The oscillogram of Fig. 16 was taken with the tube under-excited. As a result, the plate voltage swing was not very large, $e_{p\ min}$ being 1280 v. This caused the low efficiency shown. The screen potential was 500; this condition was favorable for an excess of secondary emission over the number of primary electrons captured by the screen and the screen current was negative as the oscillogram shows. On

[8] J. C. Warner, "Some Characteristics and Applications of Four-Electrode Tubes," Proc. I. R. E., 16, 424; April, 1928.

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account of the screening effect, plate current wave shape was determined almost entirely by the positive grid voltage loop, the variation in plate voltage having very little effect. Since the positive grid voltage loop was not noticeably distorted from sinusoidal shape, the plate current looks like a section of a sine wave. The plate current leads the plate voltage slightly. This is because the frequency was always adjusted for minimum plate current. Under this condition the tank circuit has a capacitative reactance, but nevertheless, maximum efficiency is obtained.

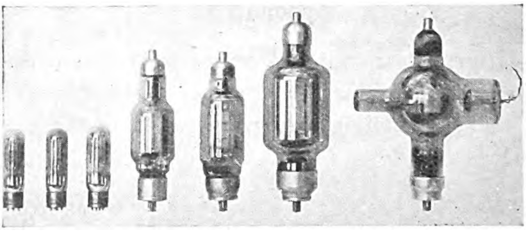

Fig. 18—Tubes Used in This Work.
Left to right: UV211, UV203A, UV845, UV204A, UV849, UV851, UV861.

Fig. 17 shows the conditions which obtained at full load and full excitation. $e_{p\,\min}$ was now 630 volts, which was sufficient to pass only 1 ampere plate current. Hence, there is a saddle in the plate current wave. The plate was also robbed of current to some extent by the screen grid and control grid. As long as e_p was well above screen potential the screen current was negative, but as e_p approached the screen voltage, conditions became unfavorable for secondary emission from the screen so the screen current reversed.

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Appendix III

AIR COOLED TRANSMITTING TUBES; THORIATED TUNGSTEN FILAMENTS

Filament Data	UV845	UV211	UV203A	UV204A	UV849	UV851	UV861
Volts	10.	10.	10.	11.	11.	11.	11.
Amperes	3.25	3.25	3.25	3.85	5.00	15.5	10.0
Electron Emission (amperes) Approx.	3.2	3.2	3.2	4.9	5.5	20.	11.
R-F Power Amplifier Use							
Maximum operating plate volts							
Modulated d-c	1000	1000	1000	2000	2000	2000	3000
Unmodulated d-c	1250	1250	1250	2500	2500	2500	4000
Maximum plate dissipation (watts)	100	100	100	250	400	750	400
Maximum d-c plate Current (amperes)	0.175	0.175	0.175	0.275	0.350	1.00	0.350
Output watts (Nominal)	75	75	75	250	350	1000	500

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

MATHEMATICAL THEORY OF THE FOUR-ELECTRODE TUBE*

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Summary—*This paper gives the mathematics of the four-electrode tube, including in the most general case expressions for the plate and two grid currents in terms of applied voltages in the two grid circuits and the impedances of all three circuits. The results are exact, variation in the amplification factors being included. Because of the complexity of the general equations, these expressions for the currents have been developed:* I, *in terms of one grid-filament voltage and external impedances in plate and other grid circuit, assuming no voltage in latter;* II, *in terms of grid-filament voltages of both grid circuits and external impedance in plate circuit;* III, *in terms of applied voltages in both grid circuits and external impedances in plate and two grid circuits.*

The results of I *show the effect of an external impedance in the non-control grid of a "screen-grid" or "space-charge-grid" tube;* II *covers the approximate theory of a "double-function" tube;* III *gives a more exact and comprehensive theory of the tube in any use.*

THE purpose of this paper is to develop general equations for the plate and grid currents of a four-electrode tube. The work is essentially an extension to the four-electrode tube of Carson and Llewellyn's work on the three-electrode tube.[1]

Let the subscript a refer to one of the two grids, b to the other; let E_p, E_a, E_b be the total instantaneous voltages between the plate, first and second grids, and the filament, respectively; and let I_p, I_a, and I_b be corresponding currents. Then if the subscript o denote steady d.c. values obtaining before the application of any a.c. voltage to the tube, and small letters indicate changes resulting from applied a.c. voltages,

$$\begin{aligned} E_p &= E_{po} + e_p \\ E_a &= E_{ao} + e_a \\ E_b &= E_{bo} + e_b \\ I_p &= I_{po} + i_p \\ I_a &= I_{ao} + i_a \\ I_b &= I_{bo} + i_b. \end{aligned} \tag{1}$$

* Dewey decimal classification: R132. Original manuscript received by the Institute, October 24, 1928.

[1] Carson, "A Theoretical Study of the Three Element Vacuum Tube," Proc. I R.E., 7, 187; April, 1919.

Llewellyn, "Operation of Thermionic Vacuum Tube Circuits," *Bell System Tech. Jour.* V, 433–462; July, 1926.

Since much of the interpretation of the results of this paper is similar to that of the latter paper, it will not be entered into in detail here.

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The amplification factors μ_a and μ_b, the plate resistance r_p, and the mutual conductances g_a and g_b are defined as follows:

$$\mu_a = \frac{\dfrac{\partial I_p}{\partial E_a}}{\dfrac{\partial I_p}{\partial E_p}} \qquad \mu_b = \frac{\dfrac{\partial I_p}{\partial E_b}}{\dfrac{\partial I_p}{\partial E_p}} \tag{2}$$

$$\frac{1}{r_p} = \frac{\partial I_p}{\partial E_p} \tag{3}$$

whence

$$\frac{\partial I_p}{\partial E_a} = \frac{\mu_a}{r_p} = g_a \qquad \frac{\partial I_p}{\partial E_b} = \frac{\mu_b}{r_p} = g_b \tag{4}$$

Note that

$$\frac{\partial g_a}{\partial E_b} = \frac{\partial g_b}{\partial E_a}.$$

It should be recalled that a partial derivative of a function of several independent variables with respect to one of them is the ordinary derivative of the function with respect to that variable when all the other variables are held constant. Hence μ_a, as defined, is to be determined for a fixed value of E_b. In addition to the above, define six analogous quantities for the grid circuits thus:

$$\nu_a = \frac{\dfrac{\partial I_a}{\partial E_p}}{\dfrac{\partial I_a}{\partial E_a}} \qquad \nu_b = \frac{\dfrac{\partial I_b}{\partial E_p}}{\dfrac{\partial I_b}{\partial E_b}} \tag{5}$$

$$\frac{1}{r_{ga}} = \frac{\partial I_a}{\partial E_a} \qquad \frac{1}{r_{gb}} = \frac{\partial I_b}{\partial E_b} \tag{6}$$

$$\frac{\partial I_a}{\partial E_p} = \frac{\nu_a}{r_{ga}} = g_{ga} \quad \text{and} \quad \frac{\partial I_b}{\partial E_p} = \frac{\nu_b}{r_{gb}} = g_{gb}. \tag{7}$$

The mutual conductances g_a and g_b defined by (4) and the reflex mutual conductances g_{ga} and g_{gb} defined by (7) stand for ratios μ_a/r_p etc., which often enter into the mathematical expressions. By Taylor's series,

$$i_p = P_1e_a + P_2e_b + P_3e_p + P_4e_a^2 + P_5e_b^2 + P_6e_p^2 + P_7e_ae_b + P_8e_ae_p + P_9e_be_p + \cdots \tag{8}$$

$$i_a = A_1e_a + A_2e_b + A_3e_p + A_4e_a^2 + A_5e_b^2 + A_6e_p^2 + A_7e_ae_b$$

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$$+A_8e_ae_p+A_9e_be_p+\cdots \tag{9}$$

$$i_b=B_1e_a+B_2e_b+B_3e_p+B_4e_a^2+B_5e_b^2+B_6e_p^2+B_7e_ae_b$$
$$+B_8e_ae_p+B_9e_be_p+\cdots \tag{10}$$

where

$$P_1=\frac{\partial I_p}{\partial E_a}=g_a \qquad P_5=\frac{1}{2}\frac{\partial^2 I_p}{\partial E_b^2}=\frac{1}{2}\frac{\partial g_b}{\partial E_b}$$

$$P_2=\frac{\partial I_p}{\partial E_b}=g_b \qquad P_6=\frac{1}{2}\frac{\partial^2 I_p}{\partial E_p^2}=-\frac{1}{2r_p}\frac{\partial r_p}{\partial E_p}$$

$$P_3=\frac{\partial I_p}{\partial E_p}=\frac{1}{r_p} \qquad P_7=\frac{\partial^2 I_p}{\partial E_a\partial E_b}=\frac{\partial g_a}{\partial E_b}=\frac{\partial g_b}{\partial E_a}$$

$$P_4=\frac{1}{2}\frac{\partial^2 I_p}{\partial E_a^2}=\frac{1}{2}\frac{\partial g_a}{\partial E_a} \qquad P_8=\frac{\partial^2 I_p}{\partial E_p\partial E_a}=\frac{\partial g_a}{\partial E_p}$$

$$P_9=\frac{\partial^2 I_p}{\partial E_p\partial E_b}=\frac{\partial g_b}{\partial E_p}$$

$$A_1=\frac{\partial I_a}{\partial E_a}=\frac{1}{r_{ga}} \qquad A_5=\frac{1}{2}\frac{\partial^2 I_a}{\partial E_b^2}=\frac{1}{2}\frac{\partial G_a}{\partial E_b}$$

$$A_2=\frac{\partial I_a}{\partial E_b}=G_a \qquad A_6=\frac{1}{2}\frac{\partial^2 I_a}{\partial E_p^2}=\frac{1}{2}\frac{\partial g_{ga}}{\partial E_p}$$

$$A_3=\frac{\partial I_a}{\partial E_p}=g_{ga} \qquad A_7=\frac{\partial^2 I_a}{\partial E_b\partial E_a}=-\frac{1}{r_{ga}^2}\frac{\partial r_{ga}}{\partial E_b}$$

$$A_4=\frac{1}{2}\frac{\partial^2 I_a}{\partial E_a^2}=-\frac{1}{2r_{ga}^2}\frac{\partial r_{ga}}{\partial E_a} \qquad A_8=\frac{\partial^2 I_a}{\partial E_a\partial E_p}=\frac{\partial g_{ga}}{\partial E_a} \tag{11}$$

$$A_9=\frac{\partial^2 I_a}{\partial E_b\partial E_p}=\frac{\partial g_{ga}}{\partial E_b}$$

$$B_1=\frac{\partial I_b}{\partial E_a}=G_b \qquad B_5=\frac{1}{2}\frac{\partial^2 I_b}{\partial E_b^2}=-\frac{1}{2r_{gb}^2}\frac{\partial r_{gb}}{\partial E_b}$$

$$B_2=\frac{\partial I_b}{\partial E_b}=\frac{1}{r_{gb}} \qquad B_6=\frac{1}{2}\frac{\partial^2 I_b}{\partial E_p^2}=\frac{1}{2}\frac{\partial g_{gb}}{\partial E_p}$$

$$B_3=\frac{\partial I_b}{\partial E_p}=g_{gb} \qquad B_7=\frac{\partial^2 I_b}{\partial E_a\partial E_b}=-\frac{1}{r_{gb}^2}\frac{\partial r_{gb}}{\partial E_a}$$

$$B_4=\frac{1}{2}\frac{\partial^2 I_b}{\partial E_a^2}=\frac{1}{2}\frac{\partial G_b}{\partial E_a} \qquad B_8=\frac{\partial^2 I_b}{\partial E_a\partial E_p}=\frac{\partial g_{gb}}{\partial E_a}$$

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$$B_9 = \frac{\partial^2 I_b}{\partial E_b \partial E_p} = \frac{\partial g_{ob}}{\partial E_b}$$

In (11), all the P's, A's and B's are constants, each derivative being evaluated at the point $E_p = E_{po}$, $E_a = E_{ao}$ and $E_b = E_{bo}$. Throughout this article this is to be understood for each derivative. The quantities G_a and G_b defined above might be termed the "cross mutual conductances."

Notation: If z is an impedance, its value at a frequency $p_m/2\pi$ may be denoted $z(p_m)$.

It will be necessary to deal with each component of a voltage or current of the form

$$e_{a1} = \frac{E_{a1}}{2} \epsilon^{j(p_1 t + \theta_1)} \tag{12}$$

(where ϵ = base nat. logs) and the corresponding impedance $z(p_1)$. The conjugate component

$$\bar{e}_{a1} = \frac{E_{a1}}{2} \epsilon^{-j(p_1 t + \theta_1)} \tag{12a}$$

has a corresponding impedance $\bar{z}(p_1)$ which is the conjugate of $z(p_1)$.

I. Effect of an Impedance in the Circuit of the Second Grid of a Four-Electrode Tube

Let us first consider a restricted problem, that in which there is an applied voltage and no external impedance in one grid circuit, say the "a" grid circuit, while there are external impedances but no applied voltages in the "b" grid circuit and the plate circuit. This would correspond to the case of a signal impressed on the "a" grid circuit, the latter having negligible external impedance, while the "b" grid circuit has an appreciable impedance.

Let the generalized impedance in the plate circuit be z and let z_b be the generalized impedance of the "b" grid circuit.

Let
$$e_a = \sum E_{an} \cos (p_n t + \theta_n)$$
$$= \sum \frac{E_{an}}{2} \left[\epsilon^{j(p_n t + \theta_n)} + \epsilon^{-j(p_n t + \theta_n)} \right] \tag{13}$$
$$i_p = \sum d_{1n} e_{an} + \sum d_{2(m+n)} (e_a^2)_{(m+n)} + \cdots \tag{13a}$$
$$i_b = \sum \beta_{1n} e_{an} + \sum \beta_{2(m+n)} (e_a^2)_{(m+n)} + \cdots \tag{13b}$$

where any term, say $\sum d_{2(m+n)} (e_a^2)_{(m+n)}$ represents a sum of terms, each one of the form $E \epsilon^{j[(p_m + p_n)t + (\theta_m + \theta_n)]}$ (or conjugate form) multiplied by a d coefficient.

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Then $e_p = -\sum z i_p = -\sum d_{1n}(ze_a)_n - \sum d_{2(m+n)}(ze_a^2)_{(m+n)} + \cdots$ (14)

$$e_b = -\sum z_b i_b = -\sum \beta_{1n}(z_b e_a)_n - \sum \beta_{2(m+n)}(z_b e_a^2)_{(m+n)} + \cdots \quad (14a)$$

where each set of brackets () represents a voltage of the form $E\epsilon^{j(p_n t + \theta_n)}$ multiplied by the appropriate impedance $z(p_n)$. Substitute in (8) and (10) and equate coefficients of terms of like order and periodicity. Then

$$d_{1n} = \frac{\mu_a[r_{gb} + z_b(p_n)] - \mu_b r_{gb} z_b(p_n) G_b}{[r_p + z(p_n)][r_{gb} + z_b(p_n)] - \mu_b \nu_b z(p_n) z_b(p_n)} \quad (15a)$$

$$\beta_{1n} = \frac{r_{gb} G_b - d_{1n} \nu_b z(p_n)}{r_{gb} + z_b(p_n)} \quad (15b)$$

$$d_{2(m+n)} = \frac{\phi r_p [r_{gb} + z_b(p_m + p_n)] - \psi \mu_b r_{gb} z_b(p_m + p_n)}{2[r_p + z(p_m + p_n)][r_{gb} + z_b(p_m + p_n)] - \mu_b \nu_b z(p_m + p_n) z_b(p_m + p_n)} \quad (15c)$$

where

$$\phi = \frac{\partial g_a}{\partial E_a} + \frac{\partial g_b}{\partial E_b}[\beta_{1n}\beta_{1m} z_b(p_n) z_b(p_m)] - \frac{1}{r_p^2}\frac{\partial r_p}{\partial E_p}[d_{1n} d_{1m} z(p_m) z(p_n)]$$
$$- \frac{\partial g_a}{\partial E_b}[\beta_{1n} z_b(p_n) + \beta_{1m} z_b(p_m)] - \frac{\partial g_a}{\partial E_p}[d_{1n} z(p_n) + d_{1m} z(p_m)]$$
$$+ \frac{\partial g_b}{\partial E_p}[d_{1n}\beta_{1m} z(p_n) z_b(p_m) + d_{1m}\beta_{1n} z(p_m) z_b(p_n)]$$

and

$$\psi = \frac{\partial G_b}{\partial E_a} - \frac{1}{r_{gb}^2}\frac{\partial r_{gb}}{\partial E_b}[\beta_{1n}\beta_{1m} z_b(p_n) z_b(p_m)] + \frac{\partial g_{gb}}{\partial E_p}[d_{1n} d_{1m} z(p_n) z(p_m)]$$
$$+ \frac{1}{r_{gb}^2}\frac{\partial r_{gb}}{\partial E_a}[\beta_{1n} z_b(p_n) + \beta_{1m} z_b(p_m)] - \frac{\partial g_{gb}}{\partial E_a}[d_{1n} z(p_n) + d_{1m} z(p_m)]$$
$$+ \frac{\partial g_{gb}}{\partial E_b}[d_{1n}\beta_{1m} z(p_n) z_b(p_m) + d_{1m}\beta_{1n} z(p_m) z_b(p_n)].$$

The d and β coefficients are complex quantities, and i_p is given as a sum of complex quantities when the d's are given by (15a) and (15c).

The coefficient $d_{2(m+n)}$ represents the general coefficient of the detection or modulation terms, for it is to be understood that the term

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$\sum d_{2(m+n)}\,(e_a{}^2)_{(m+n)}$ in (13a) represents a double sum including positive and negative values of m and n, thus giving modulation terms of the sum of the impressed periodicities (p_m+p_n), detection terms of the differences of the impressed periodicities (p_m-p_n), and including double frequency and d.c. terms as special cases. Of course for, say, a modulation term of periodicity p_m-p_n, the p_n already appearing in the equations must be replaced by $-p_n$ for a d.c. term p_n must be replaced by $-p_m$, etc.

Resistance Case. If the load in the plate circuit be a resistance r, and the impedance in the second or "b" grid circuit be a resistance r_b, then (15a) reduces to

$$d_1=\frac{\mu_a(r_b+r_{gb})-\mu_b r_b r_{gb} G_b}{(r+r_p)(r_b+r_{gb})-\mu_b\nu_b r r_b} \tag{16a}$$

and (15b) and (15c) reduce to

$$\beta_1=\frac{r_{gb}G_b(r+r_p)-\mu_a\nu_b r}{(r+r_p)(r_b+r_{gb})-\mu_b\nu_b r r_b} \tag{16b}$$

$$d_2=\frac{\begin{aligned}&r_p(r_b+r_{gb})\left[\frac{1}{2}\frac{\partial g_a}{\partial E_a}+\frac{1}{2}r_b{}^2\beta_1{}^2\frac{\partial g_b}{\partial E_b}-\frac{r^2d_1{}^2}{2r_p{}^2}\frac{\partial r_p}{\partial E_p}-r_b\beta_1\frac{\partial g_a}{\partial E_b}\right.\\&\left.-rd_1\frac{\partial g_a}{\partial E_p}+rr_bd_1\beta_1\frac{\partial g_b}{\partial E_p}\right]-\mu_b r_b r_{gb}\left[\frac{1}{2}\frac{\partial G_b}{\partial E_a}-\frac{1}{2}\frac{r_b{}^2\beta_1{}^2}{r_{gb}{}^2}\frac{\partial r_{gb}}{\partial E_b}\right.\\&\left.+\frac{r^2d_1{}^2}{2}\frac{\partial g_{gb}}{\partial E_p}+\frac{r_b\beta_1}{r_{gb}{}^2}\frac{\partial r_{gb}}{\partial E_a}-rd_1\frac{\partial g_{gb}}{\partial E_a}+rr_bd_1\beta_1\frac{\partial g_{gb}}{\partial E_b}\right]\end{aligned}}{(r+r_p)(r_b+r_{gb})-\mu_b\nu_b r r_b} \tag{16c}$$

It is evident that i_p depends in a complicated way on r_b. From d_1, conditions for maximum amplification current can be obtained by differentiation. For all cases the amplification voltage is maximum for $r=\infty$.

The coefficients reduce, when there is no second grid, to $d_1=\mu_a/r+r_p$ and

$$d_2=\frac{\left[-\mu_a{}^2r_p\frac{\partial r_p}{\partial E_p}+\mu_a(r_p{}^2-r^2)\frac{\partial\mu_a}{\partial E_p}+(r+r_p)^2\frac{\partial\mu_a}{\partial E_a}\right]}{2(r+r_p)^3} \tag{16d}$$

II. Expressions in Terms of Grid-Filament Voltages

To get the currents in terms of grid-filament voltages e_a and e_b and the external impedance z in the plate circuit, assume

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$$i_p = \sum d_{1n} e_{an} + \sum d_{2h} e_{bh} + \sum d_{3(m+n)} (e_a^2)_{(m+n)} + \sum d_{4(h+k)} (e_b^2)_{(h+k)} + \sum d_{5(n+h)} (e_a e_b)_{(n+h)} + \cdots \quad (17)$$

The subscripts n and m will be used to indicate components of e_a with periodicities p_n and p_m, respectively, while h and k will be used for components e_b, with periodicities q_h and q_k. Thus a subscript $d_{5(n+h)}$ goes with a term of frequency $(p_n + q_h)/2\pi$.

$$-e_p = \sum d_{1n} (z e_a)_n + \sum d_{2h} (z e_b)_h + \sum d_{3(m+n)} (z e_a^2)_{(m+n)} + \sum d_{4(h+k)} (z e_b^2)_{(h+k)} + \sum d_{5(n+h)} (z e_a e_b)_{(n+h)} + \cdots. \quad (17a)$$

Proceeding as before,

$$d_{1n} = \frac{g_a r_p}{r_p + z(p_n)} = \frac{\mu_a}{r_p + z(p_n)} \quad (18a)$$

$$d_{2h} = \frac{g_b r_p}{r_p + z(q_h)} = \frac{\mu_b}{r_p + z(q_h)} \quad (18b)$$

$$d_{3(m+n)} = \frac{r_p \left\{ \frac{\partial g_a}{\partial E_a} [r_p + z(p_m)][r_p + z(p_n)] - g_a^2 z(p_m) z(p_n) \frac{\partial r_p}{\partial E_p} \right\} - g_a r_p^2 \frac{\partial g_a}{\partial E_p} \{ z(p_n)[r_p + z(p_m)] + z(p_m)[r_p + z(p_n)] \}}{2[r_p + z(p_m)][r_p + z(p_n)][r_p + z(p_m + p_n)]} \quad (18c)$$

Likewise

$$d_{4(h+k)} = \frac{r_p \left\{ \frac{\partial g_b}{\partial E_b} [r_p + z(q_h)][r_p + z(q_k)] - g_b^2 z(q_h) z(q_k) \frac{\partial r_p}{\partial E_p} \right\} - g_b r_p^2 \frac{\partial g_b}{\partial E_p} \{ z(q_h)[r_p + z(q_k)] + z(q_k)[r_p + z(q_h)] \}}{2[r_p + z(q_h)][r_p + z(q_k)][r_p + z(q_h + q_k)]} \quad (18d)$$

$$d_{5(n+h)} = -\frac{r_p \left\{ g_a g_b z(p_n) z(q_h) \frac{\partial r_p}{\partial E_p} + g_b r_p z(q_h)[r_p + z(p_n)] \frac{\partial g_a}{\partial E_p} + g_a r_p z(p_n)[r_p + z(q_h)] \frac{\partial g_b}{\partial E_p} - \frac{\partial g_a}{\partial E_b} [r_p + z(p_n)][r_p + z(q_h)] \right\}}{[r_p + z(p_n)][r_p + z(q_h)][r_p + z(p_n + q_h)]} \quad (18e)$$

Each d is a complex quantity, and is the coefficient of a voltage of the form $E\epsilon^{j(pt+\theta)}$, for example, $d_{3(m+n)}$ is the coefficient of $E_{am}E_{an}/2$ $\epsilon^{+j[(pm+pn)t+(\theta m+\theta n)]}$. To change this to an expression in terms of cosines, this must be combined with its conjugate

$$\overline{d}_{3(m+n)} (\overline{e_a^2})_{(m+n)} \cdot$$

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The first two terms in i_p give

$$\sum \frac{\mu_a \cos\left[p_n t+\theta_n-\tan^{-1}\left(\dfrac{x(p_n)}{r+r_p}\right)\right]}{\sqrt{(r+r_p)^2+(x(p_n))^2}} + \sum \frac{\mu_b \cos\left[q_h t+\beta_h-\tan^{-1}\left(\dfrac{x(q_h)}{r+r_p}\right)\right]}{\sqrt{(r+r_p)^2+(x(q_h))^2}} \tag{19}$$

where $jx(p_n)$ is the reactance component of $z(p_n)$ etc.

If there is negligible external impedance in the two grid circuits, so that e_a and e_b may be taken as applied voltages and no allowance must be made for impedance drops in either external grid circuit, then the expression for i_p in terms of e_a and e_b is a complete expression for the plate current in terms of the independent variables e_a and e_b. In this case, the terms in e_a, e_b, e_a^2, and e_b^2 represent the action of each grid acting as though the other were absent; the term in e_ae_b represents a joint effect of the two grids. Thus if single frequency voltages were impressed on both grids, the modulated or detected output would appear in this term.

The modulus of $d_{3(m+n)}$ is given by a fraction, the numerator of which is

$$\frac{r_p}{2}\sqrt{\begin{aligned}&\left\{\frac{\partial g_a}{\partial E_a}[(r+r_p)^2-x(p_m)x(p_n)]-g_a^2\frac{\partial r_p}{\partial E_p}[r^2-x(p_m)x(p_n)]\right.\\ &\left.-2g_ar_p\frac{\partial g_a}{\partial E_p}[r(r+r_p)-x(p_m)x(p_n)]\right\}^2+\left\{\frac{\partial g_a}{\partial E_a}(r+r_p)[x(p_m)+x(p_n)]\right.\\ &\left.-rg_a^2\frac{\partial r_p}{\partial E_p}[x(p_m)+x(p_n)]-g_ar_p\frac{\partial g_a}{\partial E_p}(2r+r_p)[x(p_m)+x(p_n)]\right\}^2\end{aligned}}$$

and the denominator of which is

$$\sqrt{[(r+r_p)^2+x(p_m+p_n)^2][(r+r_p)^2+x(p_m)^2][(r+r_p)^2+x(p_n)^2]} \tag{20}$$

The phase angle of $d_{3(m+n)}$ is the inverse tangent of the second $\{\ \}$ over the first $\{\ \}$ minus

$$\left[\tan^{-1}\frac{x(p_m+p_n)}{r+r_p}+\tan^{-1}\frac{x(p_m)}{r+r_p}+\tan^{-1}\frac{x(p_n)}{r+r_p}\right].$$

In the case of a pure resistance load, (20) reduces to

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$$\frac{\frac{r_p}{2}\left[\frac{\partial g_a}{\partial E_a}(r+r_p)^2 - g_a{}^2 r^2 \frac{\partial r_p}{\partial E_p} - 2g_a r r_p (r+r_p)\frac{\partial g_a}{\partial E_p}\right]}{(r+r_p)^3} \tag{20a}$$

Here $g_a = \mu_a/r_p$ and if it is assumed that μ_a is constant and $\partial r_p/\partial E_a = \mu_a(\partial r_p/\partial E_p)$, then this reduces to

$$-\frac{\frac{1}{2}\mu_a{}^2 r_p \frac{\partial r_p}{\partial E_p}}{(r+r_p)^3}$$

which is the familiar coefficient for the $e_a{}^2$ term in the three-electrode tube plate current expression. (20a) and (16d) are equal.

The coefficient of $e_b{}^2(d_{4(h+k)})$ is similar to $d_{3(m+n)}$, with a, n, m, p_n and p_m replaced by b, h, k, q_h and q_k, respectively. The coefficient $d_{5(n+h)}$ is of more interest, since the $e_a e_b$ term represents the joint effect of the two grids on the plate current. The modulus of $d_{5(n+h)}$ is given by a fraction, the numerator of which is

$$r_p\sqrt{\left\{\frac{\partial g_a}{\partial E_b}[(r+r_p)^2 - x(p_n)x(q_h)] - g_a g_b \frac{\partial r_p}{\partial E_p}[r^2 - x(p_n)x(q_h)] - r_p\frac{\partial(g_a g_b)}{\partial E_p}[r(r+r_p) - x(p_n)x(q_h)]\right\}^2 + \left\{\frac{\partial g_a}{\partial E_b}(r+r_p)[x(p_n)+x(q_h)] - rg_a g_b\frac{\partial r_p}{\partial E_p}[x(p_m)+x(q_h)] - rr_p\frac{\partial(g_a g_b)}{\partial E_p}[x(p_n)+x(q_h)] - r_p{}^2\left[x(q_h)g_b\frac{\partial g_a}{\partial E_p} + x(p_n)g_a\frac{\partial g_b}{\partial E_p}\right]\right\}^2}$$

and the denominator of which is

$$\sqrt{[(r+r_p)^2 + x(p_n+q_h)^2][(r+r_p)^2 + x(p_n)^2][(r+r_p)^2 + x(q_h)^2]} \tag{21}$$

The phase angle of $d_{5(n+h)}$ is given by the inverse tangent of the second { } over the first { } minus

$$\left[\tan^{-1}\frac{x(p_n+q_h)}{r+r_p} + \tan^{-1}\frac{x(p_n)}{r+r_p} + \tan^{-1}\frac{x(q_h)}{r+r_p}\right].$$

For a pure resistance load, (21) reduces to

$$\frac{r_p\left[\frac{\partial g_a}{\partial E_b}(r+r_p)^2 - r^2 g_a g_b\frac{\partial r_p}{\partial E_p} - rr_p(r+r_p)\frac{\partial(g_a g_b)}{\partial E_p}\right]}{(r+r_p)^3} \tag{21a}$$

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and if it is further assumed that μ_a and μ_b are constants, (21a) reduces to

$$-\frac{\mu_a\mu_b r_p \dfrac{\partial r_p}{\partial E_p}}{(r+r_p)^3}. \tag{21b}$$

Since $\partial g_a/\partial E_b = \partial g_b/\partial E_a$ (see (4)), equations (21) and (21a) do not actually lack symmetry in the a's and b's.

To get i_a and i_b in terms of e_a and e_b, assume i_a and i_b are represented by series similar to (17) with d replaced by α and β, respectively.

Evaluate coefficients as before. There results

$$\begin{aligned} \alpha_{1n} &= \frac{r_p+z(p_n)(1-\mu_a\nu_a)}{r_{ga}[r_p+z(p_n)]} \\ \alpha_{2h} &= \frac{r_pG_a+z(q_h)(G_a-\mu_b g_{ga})}{[r_p+z(p_n)]} \end{aligned} \tag{22}$$

$$\alpha_{3(m+n)} = -g_{ga}z(p_m+p_n)d_{3(m+n)} - \frac{1}{2r_{ga}^2}\frac{\partial r_{ga}}{\partial E_a}$$

$$+\frac{\mu_a^2 z(p_n)z(p_m)\dfrac{\partial g_{ga}}{\partial E_p} - \mu_a\dfrac{\partial g_{ga}}{\partial E_a}\{z(p_n)[r_p+z(p_m)]+z(p_m)[r_p+z(p_n)]\}}{2[r_p+z(p_n)][r_p+z(p_m)]}$$

$$\alpha_{4(h+k)} = -g_{ga}z(q_h+q_k)d_{4(h+k)} + \frac{1}{2}\frac{\partial G_a}{\partial E_b}$$

$$+\frac{\mu_b^2 z(q_h)z(q_k)\dfrac{\partial g_{ga}}{\partial E_p} - \mu_b\dfrac{\partial g_{ga}}{\partial E_b}\{z(q_h)[r_p+z(q_k)]+z(q_k)[r_p+z(q_h)]\}}{2[r_p+z(q_h)][r_p+z(q_k)]}$$

$$\alpha_{5(n+h)} = -g_{ga}z(p_n+q_h)d_{5(n+h)} - \frac{1}{r_{ga}^2}\frac{\partial r_{ga}}{\partial E_b}$$

$$+\frac{\dfrac{1}{2}\mu_a\mu_b z(p_n)z(q_h)\dfrac{\partial g_{ga}}{\partial E_p} - \mu_b\dfrac{\partial g_{ga}}{\partial E_a}z(q_h)[r_p+z(p_n)] - \mu_a\dfrac{\partial g_{ga}}{\partial E_b}z(p_n)[r_p+z(q_h)]}{[r_p+z(p_n)][r_p+z(q_h)]}.$$

For a pure resistance load these become

$$\alpha_1 = \frac{r_p+r(1-\mu_a\nu_a)}{r_{ga}(r+r_p)}$$

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$$\alpha_2=\frac{r_pG_a+r(G_a-\mu_b g_{ga})}{r+} \tag{22a}$$

$$\alpha_3=-\frac{1}{2r_{ga}^2}\frac{\partial r_{ga}}{\partial E_a}+\frac{\mu_a^2r^2\frac{\partial g_{ga}}{\partial E_p}-2\mu_a r(r+r_p)\frac{\partial g_{ga}}{\partial E_a}}{2(r+r_p)^2}$$
$$-\frac{g_{ga}rr_p\left[(r+r_p)^2\frac{\partial g_a}{\partial E_a}-r^2g_a^2\frac{\partial r_p}{\partial E_p}-2\mu_a r(r+r_p)\frac{\partial g_a}{\partial E_p}\right]}{2(r+r_p)^3}$$

$$\alpha_4=\frac{1}{2}\frac{\partial G_a}{\partial E_b}+\frac{\mu_b^2r^2\frac{\partial g_{ga}}{\partial E_p}-2\mu_b r(r+r_p)\frac{\partial g_{ga}}{\partial E_b}}{2(r+r_p)^2}$$
$$-\frac{g_{ga}rr_p\left[(r+r_p)^2\frac{\partial g_b}{\partial E_b}-r^2g_b^2\frac{\partial r_p}{\partial E_p}-2\mu_b r(r+r_p)\frac{\partial g_b}{\partial E_p}\right]}{2(r+r_p)^3}$$

$$\alpha_5=-\frac{1}{r_{ga}^2}\frac{\partial r_{ga}}{\partial E_b}+\frac{\mu_a\mu_b r^2\frac{\partial g_{ga}}{\partial E_p}-r(r+r_p)\left(\mu_a\frac{\partial g_{ga}}{\partial E_b}+\mu_b\frac{\partial g_{ga}}{\partial E_a}\right)}{(r+r_p)^2}$$
$$-\frac{g_{ga}rr_p\left[(r+r_p)^2\frac{\partial g_a}{\partial E_b}-r^2g_ag_b\frac{\partial r_p}{\partial E_p}-rr_p(r+r_p)\frac{\partial (g_ag_b)}{\partial E_p}\right]}{(r+r_p)^3}.$$

The β coefficients are similar, β_{2h} corresponding to α_{1n}, β_{1n} to α_{2h}, $\beta_{3(m+n)}$ to $\alpha_{4(h+k)}$, $\beta_{4(h+k)}$ to $\alpha_{3(m+n)}$ and $\beta_{5(n+h)}$ to $\alpha_{5(n+h)}$.

III. Expressions in Terms of Applied Voltages

So far the plate and grid currents have been obtained in terms of the instantaneous voltages between the grids and filament (e_a, e_b). If there is an external impedance drop in either grid circuit (or both), then it is desirable to determine the currents in terms of the applied voltages e_a' and e_b'. Such a solution for i_p in terms of e_a', e_b' and the impedances should account for the action of a four-electrode tube when used as a double function tube. It should also account for the action of the tube when used as a "screen-grid" or a "space-charge-grid" tube, if current flows in the non-control grid, for only insofar as the impedance of the non-control grid circuit may be neglected may the customary equations of the three-electrode tube be used.

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To return then to the original problem, let there be an external resistance r and reactance x in the plate circuit; and let r_a and r_b and x_a and x_b be corresponding quantities in the "a" and "b" grid circuits, respectively.

$$\text{Let} \quad e_a' = \sum E_{an} \cos(p_n t + \theta_n) \tag{23a}$$

be the applied voltage in the a grid circuit, and let

$$e_b' = \sum E_{bh} \cos(q_h t + \beta_h) \tag{23b}$$

be the applied voltage in the "b" grid circuit. Then

$$e_a = e_a' - \sum z_a i_a \tag{23c}$$

$$e_b = e_b' - \sum z_b i_b \tag{23d}$$

$$\begin{aligned}\text{Let } i_p = {} & \sum h_{1n} e_{an}' + \sum h_{2h} e_{bh}' + \sum h_{3(m+n)} (e_a'^2)_{(m+n)} \\ & + \sum h_{4(h+k)} (e_b'^2)_{(h+k)} + \sum h_{5(n+h)} (e_a' e_b')_{(n+h)} + \cdots \end{aligned} \tag{24}$$

and let i_a and i_b be similar series with h replaced by a and b respectively.

Then

$$\begin{aligned} e_a = e_a' - {} & \sum a_{1n} (z_a e_a')_n - \sum a_{2h} (z_a e_b')_h - \sum a_{3(m+n)} (z_a e_a'^2)_{(m+n)} \\ & - \sum a_{4(h+k)} (z_a e_b'^2)_{(h+k)} - \sum a_{5(n+h)} (z_a e_a' e_b')_{(n+h)} - \cdots \end{aligned} \tag{24a}$$

e_b is similar, and $e_p = -\sum z i_p$.

Hence

$$\left.\begin{aligned} h_{1n} &= d_{1n} [1 - a_{1n} z_a(p_n)] - d_{2n} b_{1n} z_b(p_n) \\ h_{2h} &= d_{2h} [1 - b_{2h} z_b(q_h)] - d_{1h} a_{2h} z_a(q_h) \\ h_{3(n+m)} &= -d_{1(n+m)} z_a(p_m + p_n) a_{3(n+m)} \\ & \quad - d_{2(n+m)} z_b(p_n + p_m) b_{3(m+n)} + d_{3(m+n)} \\ & \quad [1 - a_{1n} z_a(p_n)] [1 - a_{1m} z_a(p_m)] + d_{4(n+m)} \\ & \quad b_{1n} b_{1m} z_b(p_n) z_b(p_m) - d_{5(m+n)} [1 - a_{1m} z_a(p_m)] b_{1n} z_b(p_n) \\ & \quad - d_{5(n+m)} [1 - a_{1n} z_a(p_n)] b_{1m} z_b(p_m) \\ h_{5(n+h)} &= -d_{1(n+h)} a_{5(n+h)} z_a(p_n + q_h) - d_{2(n+h)} b_{5(n+h)} z_b(p_n + q_h) \\ & - 2 d_{3(n+h)} [1 - a_{1n} z_a(p_n)] a_{2h} z_a(q_h) - 2 d_{4(n+h)} [1 - b_{2h} z_b(q_h)] b_{1n} z_b(p_n) \\ & + d_{5(n+h)} a_{2h} b_{1n} z_a(q_h) z_b(p_n) + d_{5(n+h)} [1 - a_{1n} z_a(p_n)] [1 - b_{2h} z_b(q_h)] \end{aligned}\right\} \tag{25}$$

Here the d's have already been given in (18) and the a's are given by

$$a_{1n} = \frac{1}{\phi_n} [\alpha_{1n} + z_b(p_n)(\alpha_{1n}\beta_{2n} - \alpha_{2n}\beta_{1n})] \tag{26a}$$

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$$a_{2h}=\alpha_{2h}/\phi_h \tag{26b}$$

$$a_{3(m+n)}=\frac{1}{\phi_{(n+m)}}\{\delta_{3(n+m)}[1-a_{1n}z_a(p_n)][1-a_{1m}z_a(p_m)]$$
$$+\delta_{4(n+m)}b_{1n}b_{1m}z_b(p_n)z_b(p_m)-\delta_{5(n+m)}\{[1-a_{1n}z_a(p_n)]b_{1m}z_b(p_m)$$
$$+[1-a_{1m}z_a(p_m)]b_{1n}z_b(p_n)\} \tag{26c}$$

$$a_{4(h+k)}=\frac{1}{\phi_{(h+k)}}\{\delta_{3(h+k)}a_{2h}a_{2k}z_a(q_h)z_a(q_k)$$
$$+\delta_{4(h+k)}[1-b_{2h}z_b(q_h)][1-b_{2k}z_b(q_k)]-\delta_{5(h+k)}\{[1-b_{2h}z_b(q_k)]$$
$$a_{2k}z_a(q_k)+[1-b_{2k}z_b(q_k)]a_{2h}z_a(q_k)\} \tag{26d}$$

$$a_{5(n+h)}=\frac{-1}{\phi_{(n+h)}}\{2\delta_{3(n+h)}[1-a_{1n}z_a(p_n)]a_{2h}z_a(q_h)$$
$$+2\delta_{4(n+h)}[1-b_{2h}z_b(q_h)]b_{1n}z_b(p_n)$$
$$-\delta_{5(n+h)}\{[1-a_{1n}z_a(p_n)][1-b_{2h}z_b(q_h)]$$
$$+a_{2h}b_{1n}z_a(q_h)z_b(p_n)\} \tag{26e}$$

where

$$\phi_n=[1+\beta_{2n}z_b(p_n)][1+\alpha_{1n}z_a(p_n)]-\alpha_{2n}\beta_{1n}z_a(p_n)z_b(p_n)$$
$$\delta_{3(m+n)}=\alpha_{3(m+n)}-(\beta_{3(n+m)}\alpha_{2(n+m)}-\alpha_{3(m+n)}\beta_{2(n+m)})z_b(p_m+p_n)$$
$$\delta_{4(m+n)}=\alpha_{4(m+n)}-(\beta_{4(m+n)}\alpha_{2(m+n)}-\alpha_{4(m+n)}\beta_{2(n+m)})z_b(p_m+p_n) \quad \text{etc.}$$

The corresponding b's may be written down by comparison with the a's, b_{2h} corresponding to a_{1n}, etc.

The coefficient h_{1n} may be reduced to

$$h_{1n}=\frac{d_{1n}\left[1+\dfrac{z_b(p_n)}{r_{gb}}\right]-d_{2n}z_b(p_n)G_b}{\left[\dfrac{r_{ga}+z_a(p_n)}{r_{ga}}\right]\left[\dfrac{r_{gb}+z_b(p_n)}{r_{gb}}\right]-G_aG_bz_a(p_n)z_b(p_n)} \cdot$$
$$-d_{1n}z(p_n)z_a(p_n)\left[g_{ga}+z_b(p_n)\left(\frac{g_{ga}}{r_{gb}}-G_ag_{gb}\right)\right]$$
$$-d_{2n}z(p_n)z_b(p_n)\left[g_{gb}+z_a(p_n)\left(\frac{g_{gb}}{r_{ga}}-G_bg_{ga}\right)\right] \tag{27a}$$

For $z_a=z_b=o$, this reduces to d_{1n} of course, while for $z_a=o$, it reduces to

$$\frac{\mu_a[r_{gb}+z_b(p_n)]-\mu_b r_{gb}z_b(p_n)G_b}{[r_{gb}+z_b(p_n)][r_p+z(p_n)]-\mu_b\nu_b z(p_n)z_b(p_n)} \text{ which is (15a).}$$

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For $z_b=0$, $z_a\neq 0$, which is the general three-electrode tube case, h_{1n} reduces to

$$\left(\frac{\mu_a}{r_p+z(p_n)}\right)\left[\frac{r_{ga}}{r_{ga}+z_a(p_n)\left(1-\frac{\mu_a\nu_a z(p_n)}{r_p+z(p_n)}\right)}\right] \tag{27b}$$

which is the general form of the coefficient of the amplification term of the plate current of a three-electrode tube when grid current and external grid impedance are considered. This further reduces to the customary $\mu_a/r_p+z(p_n)$ when $r_{ga}>>z_a(p_n)$.

The coefficient h_{2h} is similar.

For $z_a=z_b=0$, $h_{3(m+n)}$ reduces to $d_{3(m+n)}$ while for $z_b=0$, $z_a\neq 0$, the case of the three-electrode tube

$$\begin{aligned} h_{3(m+n)} = &-d_{1(n+m)}z_a(p_m+p_n)a_{3(m+n)} \\ &+d_{3(m+n)}\left[1-a_{1n}z_a(p_n)\right]\left[1-a_{1m}z_a(p_m)\right]. \end{aligned} \tag{27c}$$

In this case, a_{1n} is given by

$$a_{1n}=\frac{1-\frac{\mu_a\nu_a z(p_n)}{[r_p+z(p_n)]}}{r_{ga}+z_a(p_n)\left[1-\frac{\mu_a\nu_a z(p_n)}{r_p+z(p_n)}\right]}$$

Llewellyn has used the coefficient (27c) to explain grid-leak detection in a three-electrode tube. Not all the terms depend on curvature of the plate characteristic, so that there appears in i_p, among others, a frequency $(p_m-p_n)/2\pi$, which is a detection term when p_m is a side-band and p_n the carrier periodicity, when the curvature of the plate characteristic is neglected. This detection term may be made appreciable by properly choosing z_a.

In the case of the four-electrode tube it can be seen from (25), without going into detail, that if the impedance z_b is appreciable at either radio or audio frequencies, the coefficient of the detection term will be dependent upon z_b as well as z_a when the signal (carrier and sidebands) is impressed on the "a" grid. If two signals be detected at the same time, one on each grid, there will be in addition to the usual components, others due to the term $e_a'e_b'$ which may be of the same order of magnitude as the desired detection terms.

The conditions for oscillation may be obtained from (27b) for the three-electrode tube; from (27a) for the four-electrode tube. It is desired that a current of frequency $p_n/2\pi$ appear in the plate circuit

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when there is no applied voltage in the grid circuits. In this case, since h_{1n} and h_{2h} will be multiplied by zero, the denominators must be zero, and the equating of the denominators to zero will give the conditions which must be fulfilled for the tube to produce oscillations. The tube may produce oscillations of two arbitrarily different frequencies, provided the denominator of h_{1n} can be made zero for one frequency, say p_n, while the denominator of h_{2h}, which is of exactly the same form as the denominator of h_{1n}, is made zero for the other frequency, say q_h. If this be achieved, then due to the term in $e_a'e_b'$ there should be a beat frequency—the tube will act as a beat-frequency oscillator.

The complexity of the coefficients does not warrant writing out the others here; for any specific application approximations can usually be made simplifying them considerably.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

DESIGN METHODS FOR SOFT MAGNETIC MATERIALS IN RADIO*

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Summary—*The present paper is concerned with a study of design methods for soft magnetic materials such as those used in radio broadcast receivers and similar apparatus. The design of such devices as interstaged transformers calls for more than choosing points on curves in a standard handbook and computing mechanical or electrical details by ordinary power transformer theory. Radio transformer design is usually complicated by the effect of passing d-c space current through the winding or by the use of an air-gap to prevent saturation of the core by unidirectional current. The procedure to follow when one of these disturbing effects is present is not very difficult, but it becomes somewhat involved if an efficient design is wanted when both are present. In this paper an attempt is made to present a workable method of taking account of these disturbing effects. The method involves the determination of the apparent and maximum obtainable permeabilities of the magnetic materials so that a proper conclusion may be arrived at by an application of simple engineering methods.*

Introduction

The present paper is concerned with a study of design methods for soft magnetic materials such as those used in radio broadcast receivers and similar apparatus. Such a study involves the determination of the apparent and maximum obtainable permeabilities of soft magnetic materials, then showing the application of these results to definite engineering problems.

Apparent Permeability—Air Gap in Iron Path

When there is no air gap in an iron circuit, as for example in an ordinary ring sample, we can calculate the a-c permeability $\mu_{a.c.}$ from measurments of the inductance of the coil by means of the following formula:

$$\mu_{a.c.} = \frac{7.95 L_{a.c.} l 10^7}{k_1 N^2 A} \tag{1}$$

Where:

$L_{a.c.}$ = a-c inductance of reactor in henries
l = length of magnetic path in cm
k = stacking factor of core
N = number of turns of coil
A = cross sectional area of core in sq. cm.

* Dewey decimal classification: R283. Original manuscript received by the Institute, September 5, 1928. Presented by Dr. Minton before the Toronto section of the Institute, April 18, 1928.

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For direct measurements of $\mu_{a.c.}$ it is best to place two identical windings on the ring sample and measure simultaneously currents in one and voltage in the other, using an r.m.s. vacuum-tube voltmeter. The tests should best be made at 60 cycles.

As apparatus, using soft magnetic materials, usually have a superposed direct current in radio circuits, it becomes important to measure the a-c permeability for various values of d-c magnetizing forces or flux densities in the core. Accordingly, if a third winding of a few turns is wound on the ring sample and a direct current passed through it in series with an external air core inductance of high value, it is possible to calculate the d-c magnetizing force, H_o, from the following formula

$$H_o = \frac{1.256IN}{l} \tag{2}$$

where

I = current in amperes flowing in d-c winding
N = number of turns of d-c winding
l = length of iron path

Then one is able to measure the a-c permeability under various values for the d-c magnetizing force. Also, if one has before him the magnetization curves, he will be able to determine the d-c flux density for the various values of H_o used in determining the a-c permeability.

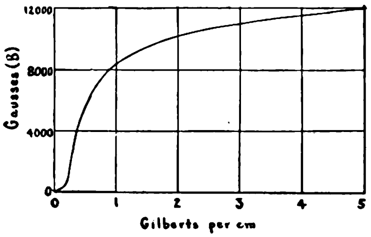

Fig. 1—Sample Soft Magnetic Alloy Magnetization Curve Used in Sample Calculations.

It will be found that the a-c permeability decreases rapidly with increasing d-c flux density. Consequently, in design work we are interested in the effective permeability of the alloys as used in actual devices. Hence, we have to consider the apparent permeability of the material where use is made of an air gap.

The introduction of the air gap in the magnetic circuit causes a decrease in the d-c flux density and a consequent rise in the apparent

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permeability. However, too large a gap cannot be used for it will tend to introduce too much reluctance to the a-c flux with a consequent lowering of the apparent permeability. In dealing with apparent permeability, then, there is a maximum obtainable value which corresponds to a definite condition with relation to the d-c magnetizing force, the air gap, and a-c flux density. If the air gap is reduced to zero, then the apparent and a-c permeabilities become identical. It becomes a matter of considerable importance, therefore, to study the introduction of an air gap in a magnetic device of the types we are considering, in order to obtain at a minimum cost the maximum effects desired. It is felt that this problem has not been given the attention it deserves, and it is hoped the following treatment will be useful.

Consider the magnetic circuit of a device, such as a transformer consisting of an iron path with a gap in it. Let l_1 and l_2 be the lengths of the iron and air paths, respectively. The magnetomotive force is consumed by the drop across the reluctance of the iron path and that across the air gap. Let H_1 and H_2 be the magnetic potential gradients along the iron and air paths respectively. The total magnetomotive force is given by:

$$(\text{m.m.f.}) = H_1 l_1 + H_2 l_2 \tag{3}$$

In the air gap the flux density is equal to the magnetizing force in the gap. If the former is represented by B_o, then $H_2 = B_o$, and when this result is substituted in (3), we get:

$$(\text{m.m.f.}) = H_1 l_1 + B_o l_2 \tag{4}$$

The two unknown quantities in (4) are H_1, the magnetizing force in the iron, and B_o, the flux density corresponding to this force. The relation between them is given by the magnetization curve. Equation (4) represents a straight line expressing the relation between H_1 and B_o. This straight line intersects the vertical axis of B at:

$$B_o = \frac{(\text{m.m.f.})}{l_2} \quad \text{when} \quad H_1 = 0 \tag{5}$$

and the horizontal axis of H at

$$H_1 = \frac{(\text{m.m.f.})}{l_1} \quad \text{when} \quad B_o = 0. \tag{6}$$

When we plot this straight line on the magnetization curve sheet, the former intersects the latter curve at some definite point. The coordinates (B_o and H_1) of this point satisfy both the magnetization

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curves and (4), and enable us to determine graphically the d-c flux density in the core under the given conditions.

If we take the derivative of H_1 with respect to B_o in (4), we get:

$$\frac{dH_1}{dB_o} = -\frac{l_2}{l_1} \cdot \tag{7}$$

That is, the slope of the straight line represented by (4) enables us to determine the ratio of the air gap to the iron path. So, when the latter is selected, the magnitude of the air gap is determined immediately.

Considering the iron path, the a-c reluctivity is the reciprocal of the a-c permeability; that is:

$$\nu_{a.c.} = \frac{1}{\mu_{a.c.}} \cdot \tag{8}$$

The reluctance R_1 of the iron path is:

$$R_1 = \nu_{a.c.}\frac{l_1}{A} \tag{9}$$

and since the reluctivity of air is unity, the reluctance, R_2, of the air path is:

$$R_2 = \frac{l_2}{A} \cdot \tag{10}$$

Hence the reluctance of the whole path is:

$$R_1 + R_2 = \frac{\nu_{a.c.}l_1 + l_2}{A} \cdot \tag{11}$$

The apparent reluctivity is then obtained by dividing (11) by the total length of the path and making A equal to unity. Hence:

$$\nu_a = \frac{\nu_{a.c.}l_1}{l_1 + l_2} + \frac{l_2}{l_1 + l_2} \cdot \tag{12}$$

In practical work the length of the gap is sufficiently small compared with the iron path to enable one to neglect l_2 in comparison with l_1 so that (12) becomes approximately:

$$\nu_a = \nu_{a.c.} + \frac{l_2}{l_1} \cdot \tag{13}$$

Hence, the apparent reluctivity is equal to the a-c reluctivity plus the ratio of the air gap to the iron path. The apparent permeability is the reciprocal of the apparent reluctivity, that is,

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$$\mu_a = \frac{1}{\nu_a} \tag{14}$$

or

$$\mu_a = \frac{l_1}{l_1 \nu_{a.c.} + l_2} \cdot \tag{15}$$

In order, therefore, to determine the apparent permeability of a soft magnetic material, it is necessary to determine beforehand the a-c reluctivity and the ratio of the air gap to the iron path. From curves it is possible to determine the a-c permeability for the alloys under any given conditions of a-c flux density, d-c flux density, and d-c magnetizing force. The reciprocal of this gives the a-c reluctivity.

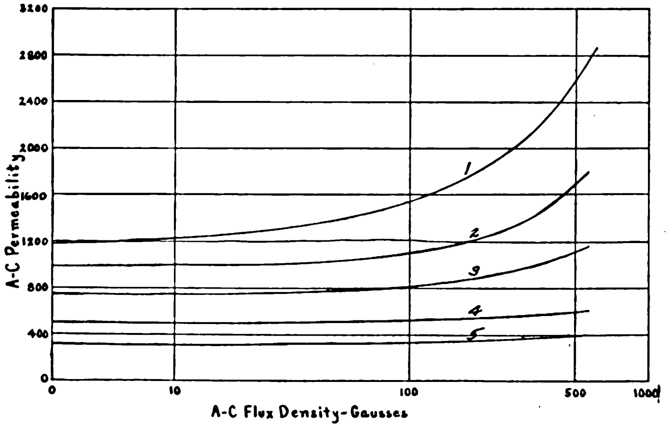

Fig. 2—Sample Soft Magnetic Alloy. A-C Permeability vs. A-C Flux Density at Various Values of D-C Flux Densities.

Curve 1—	H_o (D.C.) = 0	gilberts per cm;	B_o (D.C.)	= 0	gausses
" 2—	" = 0.5	" " "	"	= 5500	"
" 3—	" = 1.0	" " "	"	= 8400	"
" 4—	" = 2.0	" " "	"	= 10300	"
" 5—	" = 4.0	" " "	"	= 11600	"

The ratio of the air gap to the iron path can be determined graphically by means of (7), as already discussed. When this ratio and the a-c reluctivity are determined, insertion of these values in (15) [or by use of (13) and (14)] enables one to calculate the apparent permeability under the given conditions.

Maximum Obtainable Apparent Permeability

In the beginning of this discussion we referred to the maximum obtainable apparent permeability. Ordinarily, in any particular problem we have a definite direct current flowing in a winding of a definite

number of turns. This determines the d-c magnetizing force. The air gap will have its influence on the d-c and a-c flux density, as already explained, so that there is a maximum point, so to speak, where the gap can be set to secure this maximum obtainable apparent permeability under any given set of conditions.

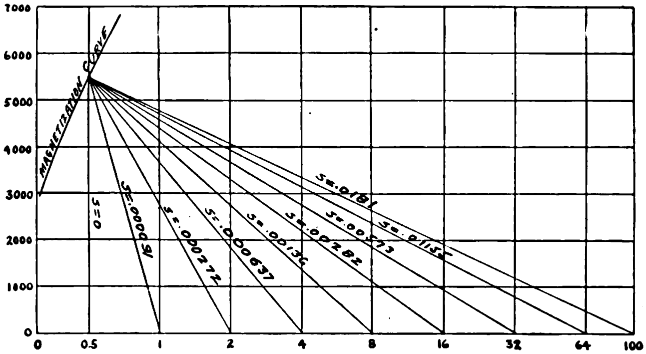

Fig. 3—Sample Soft Magnetic Alloy. Curves from which are determined the various air gaps for flux density of 5500 gausses at the various average d-c magnetizing forces.

In order to make this matter more clear, we have incorporated here the complete procedure of calculation for an alloy. The results of solutions for other alloys are given, but the details of the procedure are omitted. The procedure in all cases, however, is identical.

Calculation of Maximum Obtainable Apparent Permeability

Ordinarily we deal in practical devices with low a-c flux densities; consequently, an a-c flux density of 10 gausses has been chosen. The d-c magnetizing forces range from very small values up to perhaps 100

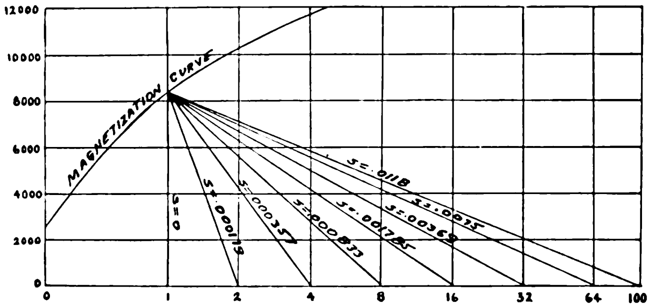

Fig. 4—Sample Soft Magnetic Alloy. Curves from which are determined the various air gaps for a flux density of 8400 gausses at the various average d-c magnetizing forces.

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gilberts or more per cm for some chokes, where an ampere or so of direct current may flow in windings of a few hundred turns. Hence, the results of our calculations extend up to a d-c magnetizing force of 100 gilberts per cm.

Our curves shown in Fig. 2 on a-c permeability are for d-c flux densities of 5500, 8400, 10,300 and 11,600 gausses. In Table I the apparent permeabilities are calculated for these d-c flux densities at various average d-c magnetizing forces up to 100 gilberts per cm. In column I of the table are found the d-c magnetizing forces. In 2 is the ratio of the air gap to the iron path l_2/l_1 at a d-c flux density of 5500 gausses.

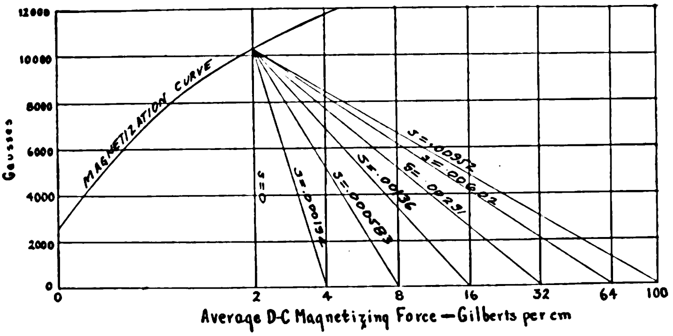

Fig. 5—Sample Soft Magnetic Alloy. Curves from which are determined the various air gaps for a flux density of 10,300 gausses at the various average d-c magnetizing forces.

These ratios are obtained by taking the slopes, in absolute value of the straight lines connecting the point, $H_0-0.5$, $B_0=5500$, on the magnetization curve to the various points on the d-c magnetizing force axis. These lines are drawn in Fig. 3. On rectangular coordinate paper the lines are straight; for convenience they are plotted on the coordinate paper with a logarithmic horizontal d-c magnetizing force axis. On this paper the lines are actually curved, so one must take the extreme ends of the lines to calculate the slopes and hence secure the ratio desired. These slopes are recorded in the 2nd, 5th, 8th, and 11th columns for the various d-c flux densities, with the corresponding curves shown in Figs. 3, 4, 5, and 6.

In accordance with (13) there must be added to these slopes the a-c reluctivity, $\nu_{a.c.}$, which is the reciprocal of the a-c permeability, $\mu_{a.c.}$, corresponding to the various d-c flux densities for an a-c flux density of 10 gausses. From curves shown in Fig. 2, the a-c permeabilities at 10 gausses a-c flux density and 5500, 8400, 10,300, and 11,600 gausses d-c flux densities are respectively 1020, 760, 520, and 325. The re-

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TABLE I

CALCULATION OF APPARENT PERMEABILITY AND RATIO OF AIR GAP

H_o Gil. per cm	$B_{a.c.}=10$, $B_o=5500$ $\mu_{a.c.}=1020$, $1/\mu_{a.c.}=0.00098$			$B_{a.c.}=10$, $B_o=8400$ $\mu_{a.c.}=760$, $1/\mu_{a.c.}=0.001316$			$B_{a.c.}=10$, $B_o=10{,}300$ $\mu_{a.c.}=520$, $1/\mu_{a.c.}=0.001925$			$B_{a.c.}=10$, $B_o=11{,}600$ $\mu_{a.c.}=325$, $1/\mu_{a.c.}=0.00308$		
	ratio of gap	ν_a	μ_a	ratio of gap	ν_a	μ_a	ratio of gap	ν_a	μ_a	ratio of gap	ν_a	μ_a
0.5	0	0.00098	1020									
1.0	0.000091	0.001071	935	0	0.001316	760						
2.0	0.000272	0.001253	800	0.000119	0.001435	687	0	0.001925	520			
4.0	0.000637	0.001617	620	0.000357	0.001673	597	0.000194	0.002119	472	0	0.00308	325
8.0	0.00136	0.00234	427	0.000833	0.002149	466	0.000583	0.002508	0.400	0.000345	0.003425	292
16.0	0.00282	0.00380	263	0.001785	0.003101	323	0.001360	0.003285	305	0.001035	0.004115	243
32.0	0.00573	0.00671	149	0.00369	0.005006	200	0.00291	0.004835	207	0.002415	0.005495	182
64.0	0.01155	0.01253	79.7	0.00750	0.008816	113.5	0.00602	0.007945	126	0.00518	0.00826	121
100.	0.01810	0.01908	52.4	0.01180	0.01312	76.2	0.00952	0.01145	87.3	0.00827	0.01135	88

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ciprocal of these values give the a-c reluctivities and are equal to 0.00098, 0.001316, 0.001925 and 0.00308. These values head the columns 3, 6, 9, and 12, Table I, corresponding to the zero air gap. The values for the apparent reluctivities are obtained by adding these a-c reluctivities to the ratio of the air gap to the iron path. These values for apparent reluctivities are given in columns 3, 6, 9, and 12. The reciprocals of these apparent reluctivities give the apparent permeabilities, as shown in columns 4, 7, 10, and 13.

We now desire the maximum obtainable apparent permeabilities corresponding to the various d-c magnetizing forces shown in the first column of Table I. For example, consider the apparent permeabilities shown in the table for 8 gilberts per cm, d-c magnetizing force. The

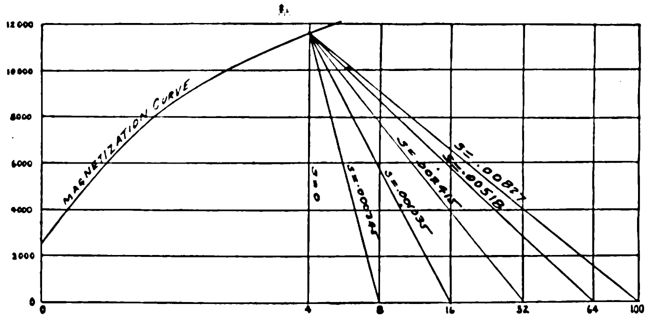

Fig. 6—Sample Soft Magnetic Alloy. Curves from which are determined the various air gaps for a flux density of 11,600 gausses at the various average d-c magnetizing forces.

values are 427, 466, 400, and 292. There is thus a maximum obtainable value of apparent permeability for 8 gilberts. The d-c flux density and the ratio of the air gap to iron path will be of some definite values to permit the attainment of this maximum value of apparent permeability.

Curves are, therefore, plotted as shown in Fig. 7. The curves are apparent permeability vs. the d-c flux density. The maximum obtainable permeability value for each curve is easily determined, and the corresponding d-c flux density is determined also. These maximum values are then plotted against the d-c magnetizing force and the resulting curve is shown in Fig. 9 curve 2.

Now, similarly, the air gap ratios shown in Table I are plotted against the apparent permeability and the air gap ratio, corresponding to the maximum obtainable apparent permeabilities, is determined graphically. See Fig. 8. These ratios corresponding to the maximum obtainable permeabilities at the various d-c magnetizing forces are plotted also in Fig. 9, curve 2.

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In an exactly similar manner the curves for other soft magnetic alloys were calculated and determined graphically. The results of this work are plotted in Fig. 9, also, so that the comparative results may be seen at once.

Application of Results

As an illustrative example of the foregoing results, let us calculate the primary inductance of an audio interstage transformer having a closed core of the various alloys discussed above. Suppose it is a second stage transformer with 0.002 ampere flowing in a primary winding consisting of 4500 turns. Let the cross section be 3/4 in. or 0.562 sq. in.

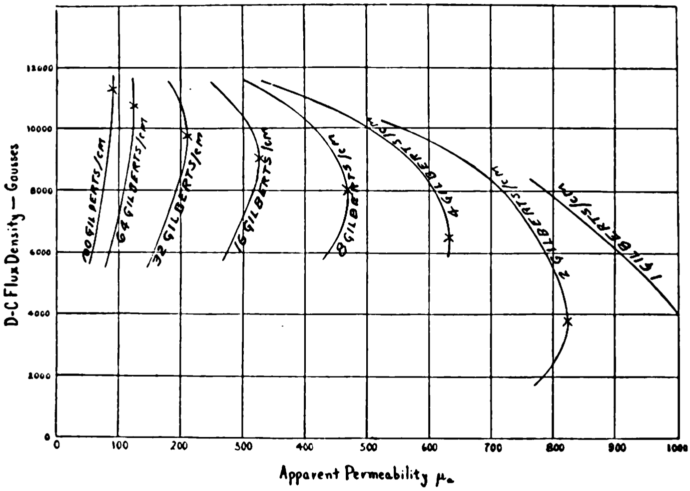

Fig. 7—Sample Alloy. Determination of maximum obtainable apparent permeability at the various d-c magnetizing forces.

Let the mean length of the iron path be 5 in. The core is to be stacked with the joints alternating side to side, so that the gap is negligible.

From (1) the a-c inductance

$$L_{a.c.} = \frac{1.256 A K_1 N^2 \mu_{a.c.}}{l} \times 10^{-8} \text{ henries.} \tag{16}$$

The cross sectional area of the iron circuit is $0.562 \times 6.45 = 3.62$ sq. cm $= A$. $K_1 = 0.875$. $N = 4500$. $l = 5 \times 2.54 = 12.7$ cm

By means of (2) the magnetizing force in gilberts per cm is

$$H_o = \frac{1.256 \times 0.002 \times 4500}{12.7} = 0.89.$$

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The a-c permeabilities for this value of d-c magnetizing force can be obtained for the various alloys by interpolating between curves nearest to this value of the magnetizing force for an a-c flux density of 10 gausses. For four different magnetic alloys these a-c permeabilities are 278, 602, 810, 1088. Substitution of these data in formula (16) gives the following results for the a-c inductance and for the reactance at 60 cycles.

Alloy	Inductance	Reactance at 60 ∿
1	17.7 henries	6,680 ohms
2	38.5	14,500
3	51.7	19,500
4	69.5	26,200

Since it is desirable to have as much voltage as possible across the primary of the transformer compared with that across the plate of the tube (say a 201A, or a 226), then the importance of selecting materials giving higher a-c permeabilities is apparent, if low frequencies are to be reproduced efficiently. This calls for using alloy No. 4.

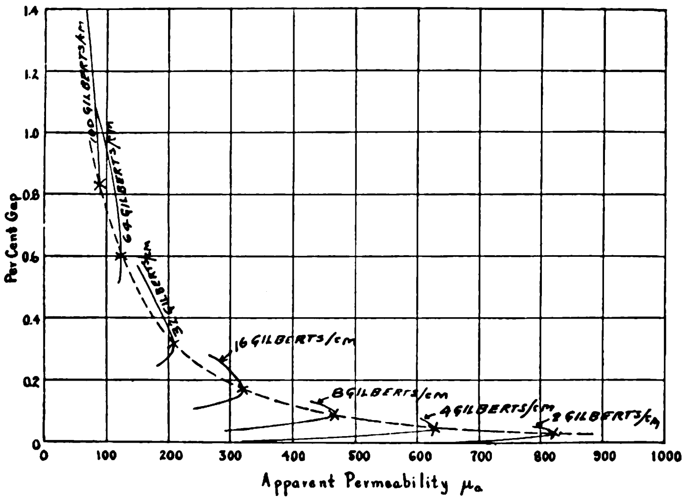

Fig. 8—Sample Alloy. Determination of air gap corresponding to maximum obtainable apparent permeability for the various d-c magnetizing forces.

As a second example let us consider an output transformer, say for a 171 tube having 0.018 ampere plate current. Let the cross sectional area of the core be 0.766 sq. in., and the mean length of the magnetic path be 6 in. Let the air gap be 0.005 in. Let us calculate the apparent inductance of the primary winding when 0.018 ampere is flowing in the primary winding and use is made of various alloys.

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The length of the magnetic path is $l = 15.2$ cm

The area of the cross section is $A = 4.95$ sq. cm

The total d-c magnetomotive force is (m.m.f.) $= 1.256\, I\, N = 90.9$ gilberts.

The average d-c magnetizing force is $H_0 -$ m.m.f.$/L = 5.95$ gilberts per cm.

The problem reduces itself to applying (13) to determine the apparent reluctivity from which we can calculate the apparent permeability. This result can be substituted in the following equation to calculate the apparent inductance

$$L_a = \frac{1.256 K_1 N^2 A_1 \mu_a}{l \times 10^8}. \tag{17}$$

This equation corresponds to (16). In (13) the second term is l_2/l_1 and this is equal to 0.0008 as given by the statement of the problem. To get $\mu_{\text{a.c.}}$ and therefore $\nu_{\text{a.c.}}$, we must calculate the d-c flux density corresponding to the given conditions. This flux is obtained

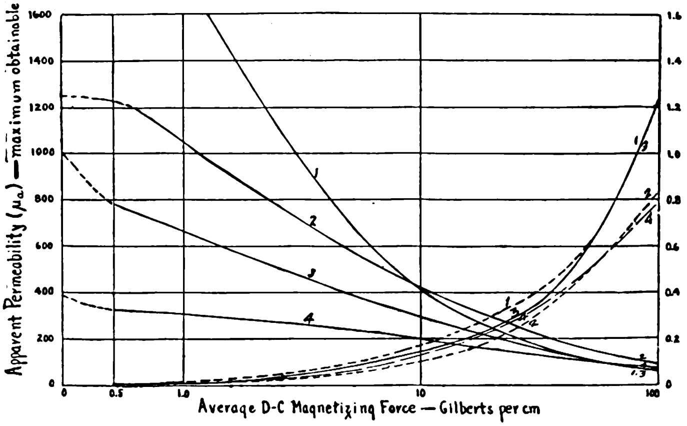

Fig. 9—Maximum obtainable apparent d-c permeability and the necessary air gap vs. d-c magnetizing force for various kinds of soft magnetic alloys.

by the intersection of the straight line, given by (4), with the magnetization curves for various alloys. The intersection of this line with the B_0 axis is, according to (5),

$$\frac{\text{m.m.m.f.}}{l_2} = \frac{90.6}{0.0127} = 7{,}140 \text{ gausses}. \tag{18}$$

The lines continue to the right with a negative slope corresponding to

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$$\frac{l_1}{l_2}=\frac{15.2}{0.0127}=1200 \text{ gausses.} \tag{19}$$

per gilbert per cm.

This line intersects the magnetization curves for the various alloys at the following d-c flux densities:

Alloy	D-C Flux Density
1	6200 gausses
2	5600 "
3	6550 "
4	5950 "

Interpolating between curves for a-c permeabilities corresponding to a low a-c flux density of 10 gausses, we obtained the following a-c permeabilities corresponding to the various d-c flux densities given above. The reciprocals of these give the a-c reluctivity.

Alloy	A-C Permeability	A-C Reluctivity
1	280	0.00357
2	525	0.001905
3	925	0.001081
4	750	0.001330

The results of the calculations for the apparent permeabilities are included in the following table:

Alloy	A-C Reluctivity	Ratio of Air Gap	Apparent Reluctivity	Apparent Permeability
1	0.00357	0.0008	0.00437	229
2	0.001905	0.0008	0.00271	369
3	0.001081	0.0008	0.00188	531
4	0.001330	0.0008	0.00213	469

From this table it is evident that the highest apparent permeability is given by using alloy No. 3, which is the best material if highest amplification is desired. These illustrations indicate how the method described in the paper is applied to particular problems and how the best solution in one case may be different from that in another.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

A NEW METHOD OF DETERMINING HEIGHT OF THE KENNELLY-HEAVISIDE LAYER*

By
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Summary—*Graphic records of radio signals transmitted from high-frequency aircraft radio transmitters are observed to show periodic variations of fairly constant frequency over considerable time intervals. A theory is evolved connecting this frequency with the transmitting distance, the ground speed of the plane and the effective height of the Kennelly-Heaviside Layer. Records of such periodic fading are reproduced from which, in accordance with this theory, the height of the layer is computed. A short discussion is given of the results obtained.*

A NUMBER of measurements have been made at the Naval Research Laboratory during the past year of field strength resulting from signals transmitted by airplanes in flight. These measurements have been taken with measuring equipment supplied by courtesy of the Bell Telephone Laboratories, which makes a graphic record on a moving chart laid off to a logarithmic scale of trans-

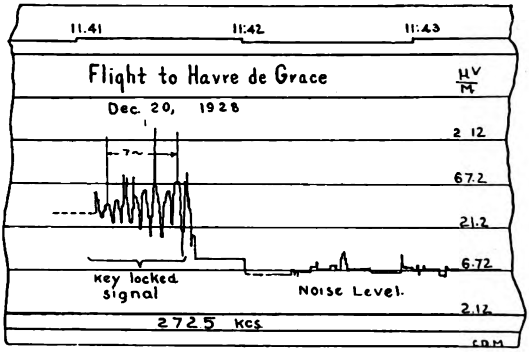

Fig. 1

mission units or TU. All of these measurements have been confined to frequencies between 2500 and 6000 kc at distances up to 130 miles (209 km). In most of them marked fading has been observed at distances beyond fifteen or twenty miles.

From data obtained during such flights which include the velocity of the plane over the ground, distance from recording station, and dis-

* Dewey decimal classification: R113.4. Original manuscript received by the Institute, February 4, 1929. Revised manuscript received April 9, 1929.

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tance between peaks of fading record, a determination of the effective height of the Kennelly-Heaviside layer has been attempted. The

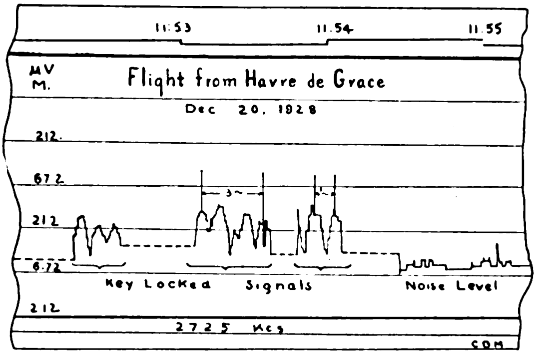

Fig. 2

method used is based upon the theory that as the plane transmitter moves from or toward the receiver, the ratio of the length of the ground path to the length of the sky path is changing, and the phase relation

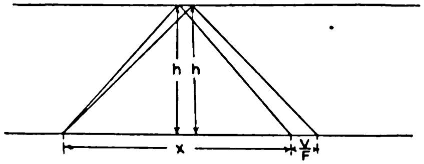

Fig. 3

between the two components is also progressively changing. When the plane has traveled the distance V/F in Fig. 3, or p–p^1 in Fig. 4, the graphic record has gone from one peak to the next and the phase of

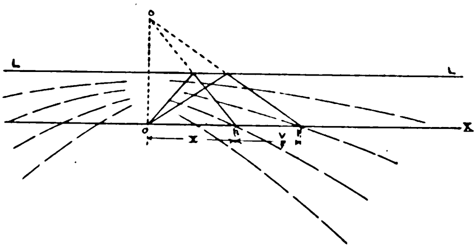

Fig. 4

one component has changed 360 deg. with respect to the other. That is, the difference between the optical lengths of the two paths has changed by a distance equal to the length of the transmitted wave.

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It is assumed that we are dealing with simple reflection from a horizontal plane at the effective height of the layer.

Let V = velocity at which transmitting plane travels over the ground in a direction to or from recording station.

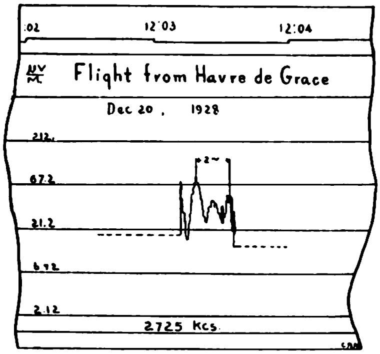

Fig. 5

τ = time between successive maxima of signal.

F = frequency of intensity variation.

X = distance from recording station.

λ = wavelength of transmission.

Then $V\tau$ (Fig. 3) $= V/F$.

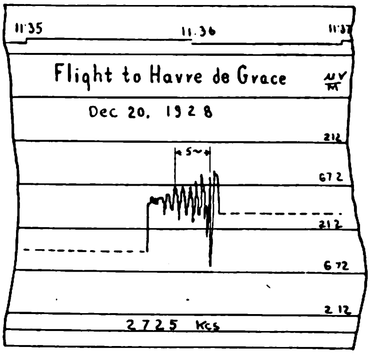

Fig. 6

From the geometry of Fig. 3

$$\sqrt{h^2+\left(\frac{x}{2}\right)^2}-\frac{x}{2}-\left[\sqrt{h^2+\left(\frac{x+V/F}{2}\right)^2}-\left(\frac{x+V/F}{2}\right)\right]=\frac{\lambda}{2} \qquad (1)$$

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Taking advantage of the fact that V/F is small compared to X, equation (1) reduces approximately to

$$h=\frac{x\sqrt{\lambda\left(\frac{2V}{F}-\lambda\right)}}{2\left(\frac{V}{F}-\lambda\right)}. \tag{2}$$

Amplifying Fig. 3 somewhat we have Fig. 4. Transmission from O reaches p, p', by both direct and reflected path. Recognizing the

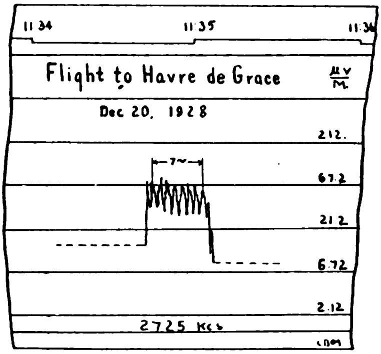

Fig. 7

similarity to well-known optical phenomena the figure is extended to indicate a similar source or image at O'. Interference patterns will

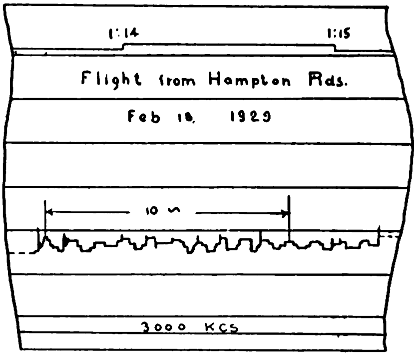

Fig. 8

then lie along hyperbolic curves with O–O' as their axis. The path of an observer along O–X will cut these lines of maximum or minimum reception.

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Thus far we have considered a fixed transmitter and a moving observer, whereas in the case of our experiment the transmitter is in the moving plane and the receiver is on the ground. However, for purpose of analysis we may consider the ground station as moving

TABLE I

Date	Time	Freq. kc f	Freq. of fading cycle sec. F	Distance between minima meters $V/F \times 10^{-2}$	Ground speed mi/hr. V	Distance of transmission miles X	Hgt. of K-H layer miles h	
12/20/28	11:35	2725	0.379	160.	135.5	48.25	73.4	*To* Havre de Grace
	11:36	2725	0.367	165.0	135.5	50.5	71.5	
	11:41	2725	0.282	215.0	135.5	60.7	53.4	
12/20/28	12:04	2725	0.1345	176.5	52.9	48.52	59.8	*From* Havre de Grace
	11:53	2725	0.1186	199.0	52.9	65.1	62.0	
2/18/29	1:19	3000	0.177	200.0	79.2	96.0	83.4	*From* Hampton Roads
	1:14	3000	0.144	246.0	79.2	103.0	70.0	
2/25/29	12:17	3000	0.243	160.0	87.6	59.65	74.0	*To* Hampton Roads
	12:48	3000	0.162	240.0	87.6	105.50	73.0	
	12:55	3000	0.170	229.0	87.6	115.25	85.0	
	1:05	3000	0.121	324.0	87.6	129.85	67.8	

relatively toward or away from the plane. With this conception the figure may correctly represent the conditions of measurement. It is significant that by this construction the intercept p–p' between fading lines grows greater as the distance between transmitter and receiver increases.

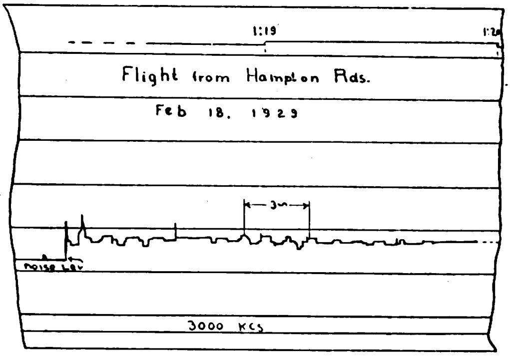

Fig. 9

A particularly good record of fading appears on sections of chart shown in Figs. 1 and 2, which were taken on December 20, 1928 between 11 A.M. and 1 P.M. at a frequency of 2725 kc. This flight was made from the U. S. Naval Air Station at Anacostia, D. C. to Havre de Grace, Md. and return over a straight course very nearly north and south.

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The altitude of transmitting plane was kept as nearly constant as possible at 2000 ft. The transmission was unmodulated CW. From position reports transmitted from the plane by radio, the ground speed was estimated at 135.5 miles per hour going and 52.9 miles per hour returning. This checks approximately with ground observations of wind velocity at 40 miles per hour. Uniform velocity in each direc-

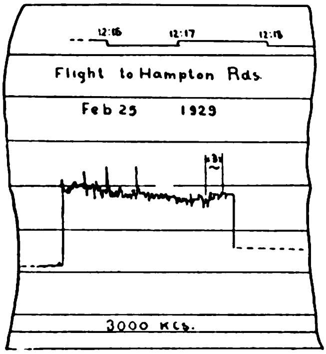

Fig. 10

tion was assumed and the position and distance of the plane for each record was computed on this basis. Frequency of fading was determined by comparing the distance between maximums on the chart with one-minute timing lines at top. The average of several cycles was taken in each case, those chosen for average being indicated on charts.

Converting to centimeters and seconds the values obtained from records 1 and 2, we have

	Outward Flight	Return Flight
X =	9,770,000 cm	10,150,000 cm
F =	0.282 cycles per sec.	0.1186 cycles per sec.
V =	6065 cm per sec.	2360 cm per sec.
λ =	11000 cm	11000 cm

Substituting these values in (2) gives the height of the reflecting layer as

Outward Flight	Return Flight
h = 54.2 miles (87.1 km)	h = 63.1 miles (101.7 km)

By a similar procedure several subsequent determinations were made as indicated by records Nos. 5 to 13 inclusive for which results are tabulated. It may be noted that in practically every case the distance between maxima of fading increases with the distance of transmission which is taken to support the theory as outlined in Fig. 4.

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The results here obtained seem in fair agreement with daytime determinations by other observers. Breit and Tuve[1] at Washington, D. C., working at a wavelength of 70 m, found a height of from 90 to

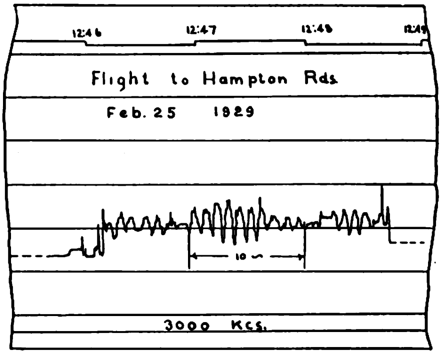

Fig. 11

220 km for the reflecting layer in daytime in summer and autumn. Taylor and Hulburt[2] calculated from observed skip distances averaged over the year for daylight conditions for waves below forty meters in length, that these waves reach heights between 120 and 240 km. Hol-

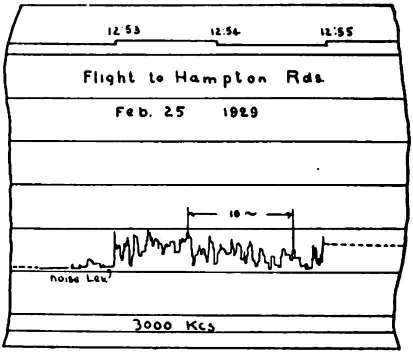

Fig. 12

lingsworth[3] in England finds 75 km in summer and 90 km in winter as the apparent daytime height reached by waves 6000 to 14,000 m long.

Observations at night show much greater heights. Appleton[4] finds indications that 400-m waves near midnight in midwinter reach heights of 300 to 400 km.

[1] Breit and Tuve, *Phys. Rev.*, **28,** 554, 1926.
[2] Taylor and Hulburt, *Phys. Rev.*, **27,** 189, 1926.
[3] Hollingsworth, *Jour.* A. I. E. E., **64,** 579, 1926.
[4] Appleton, *Nature*, **120,** 330, 1927.

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The method of this paper may be contrasted with the interference fringe or beat note method used by Appleton, Barnett, Heising and others, in that the frequency is kept constant and the distance from transmitting station is varied instead of varying the frequency and

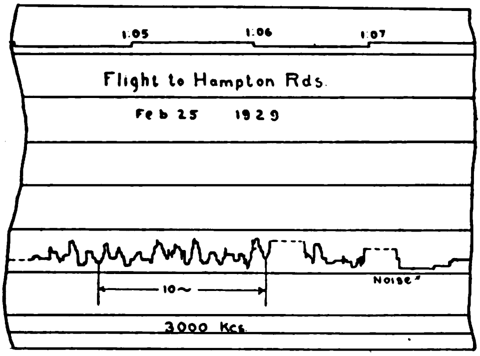

Fig. 13

observing at a constant distance from the transmitter. For this reason the observations are confined strictly to a single frequency. It is believed that these observations add one more method to those by which the effective height of the reflecting layer have been determined, and therefore contribute that much toward filling out the general theory.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

SOME OBSERVATIONS OF SHORT PERIOD RADIO FADING*

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Summary—*The data presented are the product of an investigation started at the beginning of 1928 with the object of studying the short-period fading of radio broadcast transmissions. Particular attention was paid to those intensity changes which take place during periods ranging from a few seconds to several minutes. Various antenna combinations were used in making simultaneous records in order to separate the effects of various causes of fading.*

The data secured partly confirm the conclusions of previous investigations, partly point to other sources of fading. Varying intensity of the indirect ray and interference between indirect and ground rays are evidenced as in earlier experiments, but rotation of the plane of polarization of the indirect ray is also shown to be a considerable factor and there are suggestions of lateral direction shifts of the indirect rays and of their arrival by multiple paths.

THE data presented in this paper are the product of an investigation started by the Bureau of Standards at the beginning of 1928 with the object of studying the short-period fading of radio broadcast transmissions. This study was directed particularly to those intensity changes in the receiving antenna which range, in period, from a few seconds to several minutes between peaks. These variations have been variously ascribed by previous investigators to one or more of the following causes: (1) intensity changes undergone by the indirect or atmospheric ray, namely, that portion of the radiation which, at night especially, travels to the upper atmosphere and is returned to earth at the receiving antenna by refraction; (2) interference, due to varying phase relationships of two or more indirect rays or of an indirect ray and the ground ray, which follows a direct path along the surface of the earth from transmitting to receiving antenna; (3) direction shifts, this expression being used to connote deviations of the indirect ray from the great-circle plane passing through transmitting and receiving points; (4) changes in the angle of incidence, that is, in the vertical plane of arrival of the indirect ray; (5) rotation of the plane of polarization of the atmospheric ray, this rotation being considered as a continuous process rather than in the optical sense of a

* Dewey decimal classification: R113. Original manuscript received by the Institute. Presented at meeting of International Scientific Radio Union, Brussels, Belgium, Sept. 13, 1928. Reprinted from forthcoming issue of Bureau of Standards Journal of Research. Publication approved by the Director of the Bureau of Standards of the Department of Commerce, Washington, D. C.

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definite displacement of the plane at completion of the process. The aim of the present investigation was to separate the effects due to the above-mentioned causes with a view to determining to what extent each might be a factor in fading.

Method and Apparatus

The method consisted in selecting a particular broadcasting station and making simultaneous graphic records of its carrier wave as received by different types of antennas attached to duplicate receiving sets. The antennas used were (1) a vertical single-wire, 15 meters high, with a 2-meter horizontal lead-in; (2) a vertical coil having its turns in a great-circle plane common to transmitting and receiving points, which coil will hereafter be referred to as "at maximum," or "in the

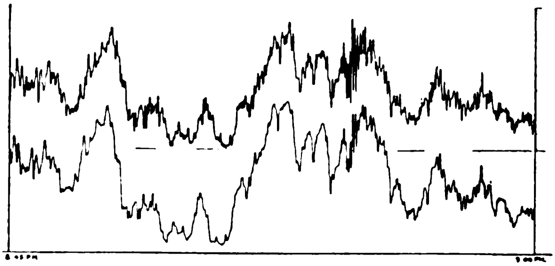

Fig. 1—WIOD (1210 kc), Miami Beach, Florida, received at Kensington, Md., 1500 km, March 15, 1928. Upper record with vertical single-wire antenna; lower record coil antenna at maximum.

maximum position;" (3) a vertical coil perpendicular to the great-circle plane, and hereafter to be designated as "at minimum," or "in the minimum position;" (4) a so-called barrage antenna, consisting of the single-wire antenna so coupled to a coil antenna in the maximum position as to introduce an out-of-phase current and thus neutralize the effect of the ground ray as well as of the vertical component of the indirect ray. The coil antennas were so balanced by a center tap arrangement as to eliminate any appreciable "antenna effect." For all tests the various antennas were located from three to five meters apart. Carefully conducted tests proved that these distances were sufficiently small so that records obtained simultaneously with similar antennas were alike, and that the distances were sufficiently large to prevent interaction between the separate receiving circuits. All observations were made at the Bureau of Standards field station located at Kensington, Maryland, 8 kilometers north of the Bureau and

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in a position practically free from "man-made" interference and distorting influences.

Other apparatus employed was essentially that described by G. W. Pickard[1] in 1923 and used later in the cooperative investigation conducted by the Bureau of Standards in 1925.[2] The receiving sets were superheterodynes of the same make and were capable of adjustment to practically the same amplification. The driving motors of the Shaw

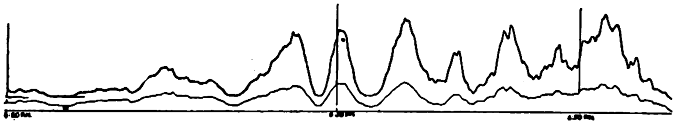

Fig. 2—WJZ (660 kc), Bound Brook, N. J., received at Kensington, Md., 300 km, May 15, 1928. Upper record, coil antenna at maximum; lower record, vertical single-wire antenna.

manual recorders were automatically synchronized every second, thus making it possible to compare two records by superposition.

Observations were concentrated largely upon two stations, but less frequent records were made on transmissions from other stations involving frequencies of 550 to 1480 kc and direct transmission paths of 13 to 1500 km. The greater number of records were made upon WJZ, Bound Brook, N. J., and upon WBAL, Baltimore, Md., because of the regularity with which each could be received, day or night, and

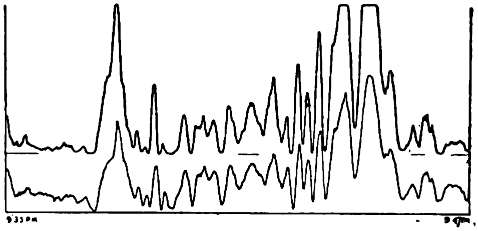

Fig. 3—WRVA (1180 kc), Richmond, Va., received at Kensington, Md., 155 km, June 1, 1928. Upper record, coil antenna at maximum; lower record, vertical single-wire antenna.

because of what appeared to be characteristic features. Many of the observations started during daylight when there was no appreciable indication of an indirect ray, and when, consequently, the most accurate bearings could be secured with the coil antennas. At this time,

[1] G. W. Pickard, "Short period variations in radio reception," Proc. I. R. E., 12, 119–158; April, 1924.

[2] J. H. Dellinger, C. B. Jolliffe, and T. Parkinson, "Cooperative measurements of radio fading in 1925," Bureau of Standards Scientific Paper No. 561, pp. 423–425.

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also, the two receiving systems could be so adjusted that the ground wave produced the same deflection on the recorder galvanometers of the two receiving systems. When records were started after dark no attempt was made to have them on an accurately comparable intensity basis, the chief aim being to study the simultaneous changes on two records rather than their exact magnitudes.

The arrangement of the pairs of simultaneous records in the illustrations will be found somewhat inconsistent since the character of the records made it easier to compare them by placing the one made with a given type of antenna sometimes above, sometimes below, the record made with another type.

Results

Simultaneous Coil and Vertical Single-Wire Antenna Records. The first method of analysis in the present investigation was by means of simultaneous fading records made with coil antenna in maximum

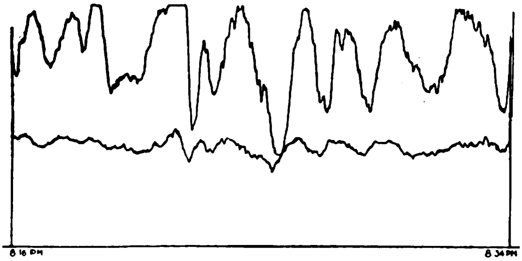

Fig. 4—WBAL (1050 kc), Baltimore, Md., received at Kensington, Md., 50 km, May 9, 1928. Upper record, coil antenna at maximum; lower record, vertical single-wire antenna.

position and with vertical single-wire antenna, respectively. Assuming both ground and reflected rays to be in the great circle plane which passes through transmitting and receiving points, certain effects were anticipated. The ground ray presumably remains normally polarized at night as during daylight; therefore, so long as there is a normally polarized component of the indirect ray any changes in the path length of the latter should result in changing phase relations between the two rays and consequently in similar and simultaneous intensity variations in the two types of antenna. Since the total electric vector in the vertical plane is effective on the coil antenna the signal strength in it should range between the sum and the difference of the two fields produced by direct and indirect rays, respectively. On the vertical single-wire antenna similar variations should be of less absolute magnitude since only the vertical component of the electric vector is effective

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and this component decreases with the angle of incidence. With either antenna the night intensity, for non-local stations at least, will sometimes exceed, sometimes fall below, the relatively constant daylight level, and similar variations should occur simultaneously on both records.

Turning now to the data and examining some fifty records involving transmission frequencies of 660 to 1480 kc and distances of 13 to 1500 km, what do we find? For stations more than about 150 km distant from the receiving antennas the results were practically as anticipated, as the typical graphic records of Figs. 1 to 3 indicate. There is a slight shifting of phase at one point in the WIOD record shown in Fig. 1 which may be real or may have been caused by changing speed of one recording drum at a time when there was difficulty

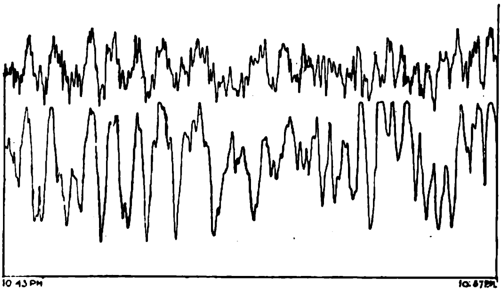

Fig. 5—WTFF (1480 kc), Mt. Vernon Hills, Va., received at Kensington, Md., 31 km, March 19, 1928. Upper record, vertical single-wire antenna; lower record, coil antenna at maximum.

with the synchronization of the two drums. Conditions prevented making another record on this station. With this single exception the intensity changes are in phase on the two antennas. An anomaly appears, however, in the records for the nearer group of stations, 13 to 53 km distant, in that the intensity changes are at times as much as 180 deg. out of phase on the two records. This holds true for variations of a few seconds duration as well as for those having periods of a minute or two. Figs. 4 to 7 show typical illustrations. Here we have the same out-of-phase effect that Appleton and Ratcliffe[3] found in simultaneous records made respectively with vertical single-wire antenna and with a loop in the maximum position, the effect of the ground ray being suppressed in the loop by means of the so-called barrage system. They accounted for the recorded differences by an interference between indirect and ground rays which was effective in

[3] E. V. Appleton and J. A. Ratcliffe, "On the nature of wireless signal variations." *Proc.* Royal Society, 115A, pp. 291–317, June 1927. See especially pp. 311–312 and Fig. 9.

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the vertical antenna but was eliminated from the loop. This explanation is not applicable to the present data since with the ground ray effective in both antennas there is still a lack of synchronism in the intensity changes of the two records. Appleton and Ratcliffe, on the other hand, found the two records varying together when the ground ray was not suppressed.[4]

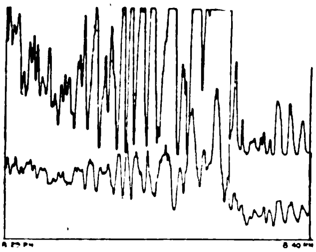

Fig. 6—WTFF (1480 kc), Mt. Vernon Hills, Va., received at Kensington, Md., 31 km, May 17, 1928. Upper record, coil antenna at maximum; lower record, vertical single-wire antenna.

A possible explanation of this anomaly may be reached by the process of elimination. Making the usual assumption that the indirect ray remains in the plane of the great circle connecting transmitter and receiver, the combined effect of ground and downcoming rays should be to cause similar changes in both vertical single-wire and coil antennas. The only exception would be when the atmospheric ray arrived

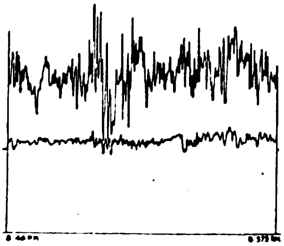

Fig. 7—WMAL, (1240 kc), Washington, D. C., received at Kensington, Md., 13 km, May 7, 1928. Upper record, coil antenna at maximum; lower record, vertical single-wire antenna.

with zero angle of incidence, in which case only the ground ray should be effective in the vertical antenna and the intensity should remain at the daylight level, while in the coil antenna the combined effect of the two rays should continue to produce fading changes. This explanation does not fit the data, since fluctuations continue on both records. It is the variation in phase relation between the two that must be ac-

[4] Reference 3, p. 300, Figs. 4 and 5.

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counted for. Rotation of the plane of polarization of the incident ray also should produce like effects in the two receiving systems; therefore, by itself, this offers no solution.

The answer is presumably associated, directly or indirectly, with the nearness of the transmitting and receiving stations. For the distances concerned, 13 to 53 km, it is perfectly possible to have direction shifts of the incident ray as great as 90 deg., due to irregularities in the refracting or reflecting upper atmosphere. This could account for changes on the vertical antenna not occurring on the coil antenna, but not for the considerable time lag between similar changes produced in the two receiving systems. If, however, there should be a lateral direction shift combined with a rotation of the plane of polarization for the downcoming ray, we should expect just such differences in the pairs of records as appear. Suppose, for purposes of illustration, that the reflected ray is arriving with constant intensity and normally

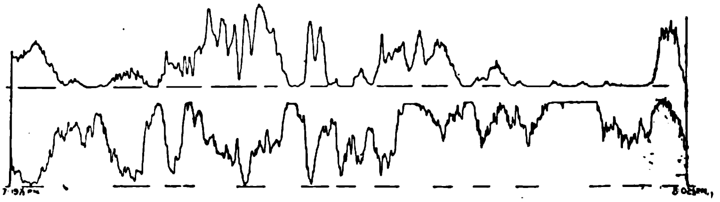

Fig. 8—WJZ (660 kc), Bound Brook, N. J., received at Kensington, Md., 300 km, April 2, 1928. Upper record, coil antenna at maximum; lower record, coil antenna at minimum position.

polarized. If, now, the lateral angle of arrival shifts until it becomes 90 deg. the signal intensity in the vertical antenna will remain constant while that in the coil antenna will decrease to a minimum, in fact, to zero, if there should be no ground wave present.

If, next, the plane of polarization rotates, the vertical component of the electric field will decrease until, with the electric field horizontal, the ray ceases to produce a current in the vertical antenna. Simultaneously, the component of the magnetic field in the vertical plane will have increased from zero to a maximum, thereby producing maximum current in the coil antenna for the given angle of incidence. Thus we have similar fading characteristics 180 deg. out of phase in the two antennas. If the direction of arrival of the incident ray gradually returns to the plane of the maximum coil antenna and the rotation of the plane of polarization continues, whether at regular or irregular speed, the phase difference between the two records will gradually decrease until it disappears. With the direction shifts more or less erratic we should have just the sort of changing phase relations shown

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on the records of Figs. 4 to 7. When rotation of the plane of polarization is absent we should find the two records in phase, as in the early part of Fig. 5 or the latter part of Fig. 6.

The existence of the direction shifts necessary to the above explanation requires proof. Bearings on stations were taken at night by means of a coil antenna, and revealed apparent direction shifts which were considerable. A convincing test, however, can only be made with an

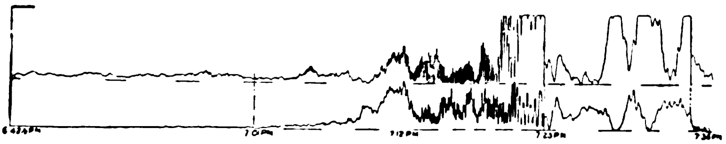

Fig. 9—WJZ (660 kc), Bound Brook, N. J., received at Kensington, Md., April 5, 1928. Local sunset at 6:36. Upper record, coil antenna at maximum; lower record, coil antenna at minimum.

antenna of the Adcock type[5] if true directions of arrival of the reflected ray are to be found. Such a test awaits the construction of further apparatus.

Interference between the indirect ray and the ground ray is evidenced on many of the records made with the coil antenna in maximum position, and to a lesser degree by those made with the vertical single-wire antenna. Wherever the intensity of one of these records falls below the daylight level the indirect ray must be to some extent out of phase

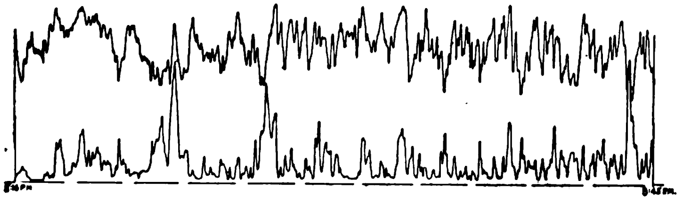

Fig. 10—WBAL (1050 kc), Baltimore, Md., received at Kensington, Md., 50 km, April 17, 1928. Upper record, coil antenna at maximum; lower record, coil antenna at minimum.

with the waves arriving along the surface of the earth. Figs. 4 to 7 give clear indications of this. Note particularly that even for stations as near as 13 km from the observing point, as shown by Fig. 7, the intensity of the downcoming ray is sometimes nearly as great as that of the ground wave, so that when the two are opposite in phase the intensity of the ground ray is largely neutralized and the record drops

[5] Improvement in means for determining the direction of a distant source of electromagnetic radiation. British Patent 130490, 1919. For actual applications see reference 14 in Appendix.

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almost to zero. Presumably in these same records the peaks are due to the addition of the reception by direct and indirect paths. It will later be shown, however, that peaks may be due to other causes, whether the ground wave be present or not.

Simultaneous Fading Records with Two Coil Antennas, at Maximum and Minimum Positions, Respectively. The second method of analysis was by simultaneous fading records of the same transmission as received by two coil antennas one in the maximum and the other in the minimum position. The purpose of this arrangement was to secure some data as to fading caused by rotation of the plane of polarization of the indirect ray; for such a phenomenon, if present, should cause an increase of intensity in either coil antenna to be accompanied by a decrease in the other. So far as possible the minimum bearing was found

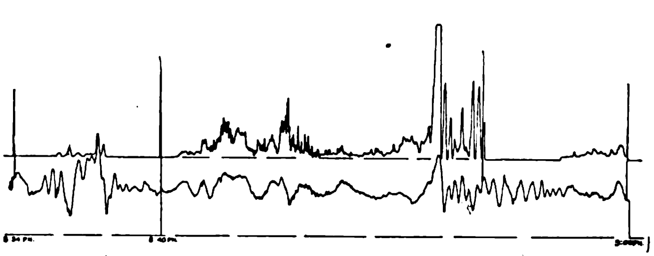

Fig. 11—WMAL (1240 kc), Washington, D. C., received at Kensington, Md., 13 km, June 5, 1928. Upper record, coil antenna at minimum; lower record, coil antenna at maximum.

for the various stations during daylight when there was no evidence of an indirect ray, and these bearings checked satisfactorily on different days. For the more distant stations, however, no ground wave could be received in daylight and it was necessary to trust to approximate bearings, which were sufficient for the project in hand.

The data secured offers considerable striking evidence of the above-mentioned phenomenon. Fig. 8[6] is a typical section cut from a pair of simultaneous records. Here it is possible to see at a glance that the two are practically 180 deg. out of phase most of the time. The minima do not always fall to zero as would be expected with the ground wave practically negligible as it is with the stage of amplification used in this case, and it therefore becomes necessary to postulate more than one indirect ray. It may be that in this record we have the effect of two rays rotating at different speeds so that their minima do not synchronize most of the time. Interference between the two is also

[6] Figs. 8 and 9 and other data indicating rotation of the plane of polarization of radio waves were shown by the author at the meeting of the American Section, U. R. S. I., in Washington, D. C., April 19, 1928.

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probably present but the effect cannot be isolated in the record. Fig. 9 is another record made on reception from the same station as Fig. 8, and is typical of what occurs during the sunset period. For approximately one-half hour after sunset the weak ray from the upper atmosphere appears to remain normally polarized and so produces intensity changes only in the coil at maximum position, possibly through a combination of interference effects and varying absorption. Evidence then gradually appears that an indirect ray somewhat abnormally polarized is producing a current in the minimum coil. With no further rotation of the plane of polarization any variations in the intensity of the indirect ray may now result in parallel changes such as appear in the first large intensity peak at 7:12. Here, due probably to the combined effects of very abnormal polarization and of small angle

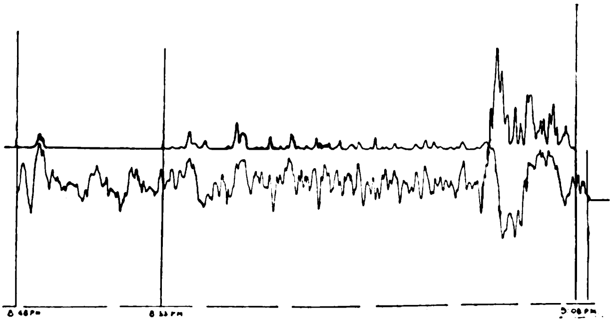

Fig. 12—WMAL (1240 kc), Washington, D. C., received at Kensington, Md., 13 km, June 6, 1928. Upper record, coil antenna at minimum; lower record, coil antenna at maximum.

of incidence accompanying increased height of the ionized layer, the intensity in the coil at minimum is practically the same as in that at maximum. From this peak on, nearly to the end of the record, the average trend of the maximum coil antenna record is 180 deg. out of phase with the minimum, indicating that the rotation of the plane of polarization of the indirect ray has gotten under way. Starting at about 7: 04 P.M. on this same record there is a superposed rapid variation having a period of from 3 to 30 seconds. This characteristic is also exactly out of phase on the two records and so suggestive of a second and relatively much weaker ray with plane of polarization rotating rapidly. Assuming the polarization to be plane, a complete rotation would have to occur in six seconds to produce the most rapid of these changes.

While these and other data seemed to be consistent with rotation of the plane of polarization of the indirect rays, there were times when

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the phase difference between the two records varied widely, sometimes becoming almost zero, as shown in Figs. 10 to 12. The ground ray could in no way be held responsible, for its effect had been eliminated by the coil antenna in minimum position, and yet variations in one antenna led or lagged behind similar variations in the other. In a recent article Eckersley[7] reports that records of high-frequency transmissions received simultaneously with vertical and horizontal antennas are opposite in phase. It is interesting to note that the illustration given in that paper, however, shows clearly the same sort of phase shifts between the two records as appear in the data of the present paper. To account for this phenomenon by an interference effect, at least two elliptically polarized indirect rays would be necessary, with their vertically polarized components undergoing changes in phase relationship similar to but lagging or leading those of the horizontally polarized components.

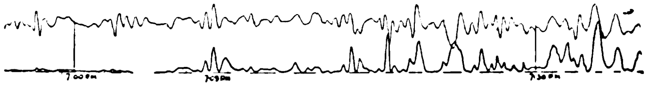

Fig. 13—WBAL (1050 kc), Baltimore, Md., received at Kensington, Md., 50 km, April 27, 1928. Local sunset at 6:58. Upper record, coil antenna at maximum; lower record, barrage minimum.

A more plausible explanation is that suggested for similar out-of-phase relations which occurred in the records made simultaneously on coil and vertical-wire antennas. If an indirect ray with rotating plane of polarization were arriving with an angle of incidence of nearly 90 deg. and the plane of its arrival should gradually shift laterally from 0 deg. to 45 deg. with respect to the maximum position of the coil antenna the records made simultaneously by coil antennas in maximum and minimum positions would gradually shift from the 180 deg. out-of-phase relationship to a nearly in-phase relationship. If the angle of incidence were considerably less than 90 deg. similar changes in phase relationships would take place as the direction of arrival shifted to the 45 deg. position, except that the two records would never come completely into phase. As the angle of incidence decreased still further the two records would be more and more out of phase. If the indirect rays always arrive by means of a single reflection the angle of incidence at the receiving end would increase with the distance between transmitting and receiving points and a direction shift of given angle would produce a greater phase shift in the pair of fading records for a distant than in the pair for a nearer station because the incident ray

[7] T. L. Eckersley, "Polarization and fading of short wireless waves," *Nature*, **121**, 707; May, 1928.

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for the former is more nearly horizontal. On the other hand, it should be possible for the ray from a nearer station to arrive at larger angles with the plane of the coil antenna in maximum position since lateral radiation from the distant station would have to go to improbable distances to be reflected back at large angles. Very distant transmissions, furthermore, would presumably be received almost entirely in the great circle plane passing through transmitting and receiving stations. In the case of possible multiple reflections of the same ray or of an inverted U-shaped path the situation would be altogether different. Further measurements of the apparent effective height of the ionized region according to the method of Appleton and Ratcliffe, providing a sufficient variety of distances be covered, may give some useful information, since this method should prove most accurate and consistent for the distant stations if there are actual lateral reflections present.

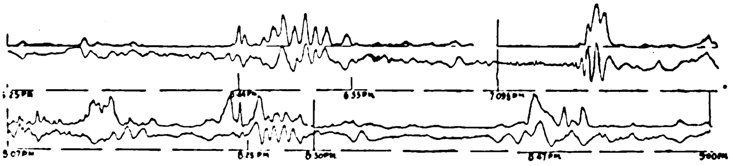

Fig. 14—WBAL (1050 kc), Baltimore, Md., received at Kensington, Md., 50 km, April 26, 1928. Local sunset at 6:57. Upper record, barrage minimum; lower record, coil antenna at maximum.

As a partial check of the possible shifts in the direction of arrival of the indirect ray, observations were made simultaneously with three receiving systems utilizing, respectively, a coil antenna in maximum position, a second in the minimum position, and a third in a vertical plane at an angle of 45 deg. with the other two. Comparisons were made vocally by the three observers as changes occurred in the indicating galvanometers, since the third recorder for securing graphic records was lacking. Observing was done at a time when the phase relation between changes in the perpendicular coils was varying. The triple observations showed that for periods of a number of minutes at a time the variations in the 45 deg. coil were more nearly in phase with those in the coil at maximum, for similar periods were more nearly in phase with changes in the coil at minimum position, and at other times were considerably out of phase with both. Such effects do not prove direction shifts to exist but are similar to what would be expected if such shifts were real.

Simultaneous Fading Records Made with Barrage Antenna and with Vertical Single-Wire or Coil Antennas. Using the barrage combination

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of antennas to eliminate effects due to the ground wave, and the coil antenna in the maximum position to receive both ground and indirect rays, it was hoped that differences in records would indicate the amount of fading due to interference between these two rays. With the indirect ray remaining in the plane of the great circle connecting the two stations, changes other than those due to interference should have similar effects on the two receiving systems, though the magnitude of changes should differ.

Figs. 13 and 14 show typical pairs of records. The earlier portion of the records of Fig. 13 occurs during sunset before the angle of incidence of the indirect ray has decreased sufficiently to produce more than a feeble horizontal component of the electric field. Consequently the intensity changes are quite feeble at first but increase as the ionized layer moves upward. During the first half of the records the intensity

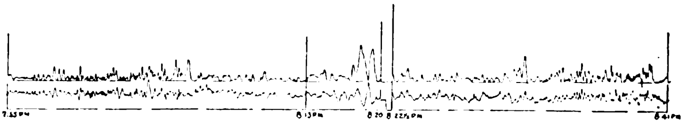

Fig. 15—WBAL (1050 kc), Baltimore, Md., received at Kensington, Md., 50 km, July 6, 1928. Upper record, barrage minimum; lower record, vertical single-wire antenna.

variations are in phase in the two antenna systems, indicating that these changes are produced farther back in the transmission path. The later portions of the records, however, show some changes in one antenna to be lagging similar changes in the other. Fig. 14 shows similar lagging effects at approximately 6:44 to 6:53, 8:25 to 8:30 and 8:47 P.M. At other times this pair of records is usually in phase or 180 deg. out. The latter condition is well explained by the fact that as the horizontal electrical component of the indirect ray increases, as shown by the barrage record, the vertical component presumably increases also, and being out of phase with the ground ray, whether partially or completely at these particular times, causes a reduction of intensity in the coil antenna set at maximum position.

Fig. 15 shows records made by a method similar to that last described except that a vertical single-wire antenna was substituted for the coil in maximum position. This method was used by Appleton and Ratcliffe.[8] Though the fading is more rapid in Fig. 15, the phase relations between the records are similar to those of Figs. 13 and 14, with less positive indications of the lagging effect because of the rapidity of variations. This effect does appear on a record[8] shown by Apple-

[8] Reference 3, pp. 311–312, and Fig. 9.

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ton and Ratcliffe, however. The explanation given by them, that the difference in records is due to the interference between indirect and ground rays which can affect only one of the two receiving systems, is a possible one for all of these pairs of records. Two facts, however, cast some question upon this as a complete solution: (1) it is doubtful if interference can account for the slight phase displacements in similar portions of the two records as it may well account for the 180 deg. out-of-phase relationships; (2) similar varying phase relationships appear in records already discussed, those made with the vertical antenna and with the coil antenna in maximum position with no elimination of the ground ray. Shifts in the direction of arrival of the downcoming ray would make the latter more effective on the vertical antenna portion of the barrage system than on either coil.

If rotating plane of polarization be combined with a direction shift in this ray the phase relationships appearing in the pairs of simultane-

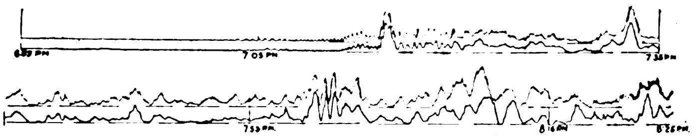

Fig. 16—WBAL (1050 kc), Baltimore, Md., received at Kensington, Md., 50 km, March 30, 1928. Local sunset at 6:30. Upper record, coil antenna at minimum; lower record, barrage minimum.

ous records become explicable. Three simultaneous records utilizing the vertical antenna, the barrage system, and the coil antenna, should throw further light on the problem.

Again, the barrage system was used simultaneously with the coil antenna in minimum position. This provided a method of eliminating results of the ground wave from both coils so that remaining effects should be due to indirect rays only and to the angles which the coils made with that indirect reception. The effect of the entire vertical electric component of this ray should, of course, be eliminated in the barrage system. Fig. 16 shows the type of record secured on transmission during the sunset period from a station only 53 km distant. The changes, both large and small, are very nearly of the same magnitude and the larger variations are usually in phase on the two records. The phase relationships of the smaller periodic variations change considerably but are most often 180 deg. out. In other words, this pair of records has much the same characteristics as appeared on the WJZ records, made with coils in maximum and minimum positions, in which the effect of the ground wave was practically absent because of

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great attenuation due to distance. There is in the early part of the records a superposed periodic variation suggestive of an indirect ray of small intensity with rotating plane of polarization and some shift in the lateral angle of arrival to cause changing phase relationships in the two fading records. The major variations would appear to be caused by a much stronger indirect ray of varying intensity with its plane of polarization unchanging at first but later rotating slowly, thus causing changes in one record to lag behind similar changes in the other. This rotation effect was not found in the records made by Appleton and Ratcliffe[9] with similar antennas. The intensity variations shown by their simultaneous records were so largely in phase that most of the fading could be explained by intensity variations occurring at some distance from the receiving antennas.

Short Period Fading. The rapid periodic fading previously mentioned and illustrated in Figs. 9 and 16 should have some significance since there is a considerable degree of regularity in its appearance, at least in reception from the two stations upon which observations have been most consistently made. During the cooperative investigation of fading conducted by the Bureau of Standards in 1925 this same phenomenon was noted,[10] though less consistently, in observations made at several different localities on transmissions from WGY (790 kc), Schenectady, N. Y. In the earlier case the phenomenon occurred for a few days at the spring equinox and was not found again. The newer cases were first noted at the same season, but have continued for several months with very few exceptions whenever records were made. On the WJZ records this type of fading starts from 15 to 30 minutes after the Washington sunset and lasts in the vicinity of a half hour. On the WBAL records, which were made much less frequently, the start is likely to be practically at sunset, though there is some variation both ways.

The explanation of this phenomenon suggested in the earlier paper as an interference effect between ground and reflected rays is now proved unsatisfactory by the fact that this periodic fading is present even when the ground ray is absent. As previously stated, the fact that such fading is out of phase on records made with coils at maximum and minimum positions, respectively, makes any interference explanation difficult and suggests a relatively weak indirect ray with rapid rotation of its plane of polarization. No explanation is given at this time for the rotation itself.

[9] Reference 3, p. 313, and Fig. 10.
[10] Reference 2, pp. 437–439 and Fig. 5, opposite p. 427.

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Conclusions

1. Considerable fading is caused by the fact that the indirect ray from a radio broadcasting station undergoes variations of intensity before arriving at the receiving antenna. This is proved by the fact that at times similar intensity changes occur simultaneously in coil antennas at maximum and minimum positions as well as in barrage antenna and coil at minimum.

2. Much evidence of fading caused by interference between ground and indirect rays is found in records of reception from transmitting stations sufficiently near to produce a ground wave at the receiving point. The night-time intensity is often less than the daytime intensity which is due to the constant ground ray. The out-of-phase indirect ray often neutralizes the effect of the ground ray partially and sometimes completely, even for short transmission paths.

3. Direction shifts are apparently necessary to the explanation of phase displacements of less than 180 deg. between otherwise similar records made simultaneously: (1) with coil antenna at maximum and with vertical single-wire antenna; (2) with coil antenna at maximum and with barrage antenna; (3) with barrage and vertical single-wire antennas; (4) with coil antennas at maximum and minimum positions; (5) with barrage and with coil antenna at minimum.

4. No proof of fading caused by fluctuating height of the ionized layer is found, but evidence of refraction of the indirect ray from a rising layer is found in the fact that the signal intensity in the coil antenna at minimum position starts at zero in daylight and gradually increases during sunset.

5. Rotation of plane of polarization of radiation refracted from the upper atmosphere is shown to be the cause of much fading, particularly, though by no means wholly, during the sunset period. The 180 deg. out-of-phase relationship between records made simultaneously with coil antennas in maximum and minimum positions seems to demand this explanation. The cause of such rotation is not explained.

6. Refractions arriving by multiple paths are evidenced in records showing a periodic type of fading superposed on the main intensity variations and yet 180 deg. out of phase on simultaneous records made with coil antennas in maximum and minimum positions. This fact makes the calculation of the effective height of the ionized layer by means of the sine of the angle of incidence, as determined by the method of Appleton and Ratcliffe,[11] somewhat doubtful, especially when measuring on other than distant stations.

[11] Reference 3, pp. 297–305.

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7. The rapid periodic type of fading noted in a previous paper as occurring in signals from one station during the sunset period has been found to be quite common in the reception from two other stations. The newer data point toward rotation of plane of polarization rather than interference as the cause.

As stated in the introduction, this investigation is far from complete. It is planned to make similar measurements over a much longer period of time to observe possible seasonal effects; to make simultaneous measurements by more than two receiving systems over a wider range of frequencies; to compare results of transmissions perpendicular to the earth's magnetic field with those parallel to it; to make direction and oscillographic observations simultaneously with the fading records; to make more quantitative measurements of these observations and compare calculated heights of the refracting layer, as determined by the angle of incidence method, on transmissions involving a wide range of distances, in order to note whether there is greater consistency in measurements on the more distant stations.

ACKNOWLEDGMENTS

Special acknowledgment is made of the assistance of H. S. Shaw of Exeter, N. H., whose contribution of the salary of a research associate made possible the carrying on of this work at the Bureau of Standards.

Acknowledgment is also made of the assistance of S. A. Buckingham, R. P. Battle, and T. R. Gilliland in preparing apparatus and making the large number of records required in this work.

Appendix

SUMMARY OF CONCLUSIONS OF PREVIOUS INVESTIGATIONS

Both Pickard, in his 1923 investigation,[12] and Appleton and Ratcliffe in their recent work[13] have given the chief credit for fading to intensity changes undergone in the path of the indirect ray. Pickard ascribes these changes to varying absorption and suggests a possible explanation, while Appleton and Ratcliffe find reason to suspect that interferences between rays travelling by multiple paths are the cause. These authors agree also in crediting interference between ground and indirect rays with a minor role in causing fading; and practically eliminate direction shifts as a factor. Appleton and Barnett,[14] as

[12] Reference 1, pp. 152–158.
[13] Reference 3, pp. 314–315.
[14] E. V. Appleton and M. A. Barnett, "On some direct evidence for downward atmospheric reflection of electric rays," *Proc.* Royal Society, A, **109**, 1925, 621–641. See specially 623 and 635–636.

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well as Smith-Rose and Barfield,[15] had previously satisfied themselves that apparent direction shifts were due chiefly to elliptical polarization of the atmospheric ray, and very little, if at all, to actual changes in lateral angle with the great circle plane through transmitting and receiving points. Pickard did not discuss changing angle of incidence nor rotation of the plane of polarization of the downcoming ray in connection with his fading data. Appleton and Ratcliffe were unable to detect any evidence that these factors were affecting fading,[16] in spite of the fact that rotation of the plane of polarization of the incident ray had been accepted as explanation for apparent direction shifts. The Bureau of Standards report[17] on cooperative measurements in 1925, while accepting varying absorption as the more general cause of fading, gave considerable weight to interference effects so far as fading of nearer stations was concerned, but no data upon the other factors were then secured.

While the present paper deals only with observations in the broadcast range of frequencies and with limited methods of analysis it is well to have in mind the conclusions reached by investigations at other frequencies and by other methods. Working particularly with frequencies above 3000 kc Breit and Tuve[18] found fading occurring independently of interference between ground and indirect rays and suggested multiple reflections and flickering effect due to wavy character of surface of the ionized region. Heising[19] as well as Taylor and Young,[20] using an oscillographic method similar to that of Breit and Tuve found proof of a multiplicity of paths of different lengths. Heising found these regularly accompanied by fading and simultaneous with an apparent rising of the ionized layer. Rukop[21] in working with transmission of pictures at high frequency had earlier secured evidences of multiple paths which were accompanied by fading which, however, was not necessarily associated with the multiplicity of paths. The path length changed at times as rapidly as 50 km a second giving the Doppler effect of a change in frequency at the

[15] R. L. Smith-Rose and R. H. Barfield, "The cause and elimination of night errors in radio direction-finding," *Jour.* I. E. E., **64**, 831–837, 1926.

[16] Reference 3, pp. 312–315.

[17] Reference 2, pp. 447–448.

[18] G. Breit and M. A. Tuve, "A test of the existence of the conducting layer," *Phys. Rev.*, **28**, 554–575, September, 1926. See especially p. 564.

[19] R. A. Heising, "Experiments and observations concerning the ionized regions of the atmosphere," Proc. I. R. E., **16**, 75; January, 1928.

[20] A. Hoyt Taylor and L. C. Young, "Studies of high-frequency radio wave propagation," Proc. I. R. E., **16**, 561; May, 1928.

[21] Von H. Rukop, "Die Bildtelegraphie als Untersuchungsmethode für die Ausbreitung der kurzen Wellen," *Elektrische Nachrichten-Technik*, **3**, 316–318, 1926.

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receiving system. From theoretical considerations Pedersen[22] concluded that there should be an infinite number of downcoming rays at a given point on the earth's surface.

Fading produced by interference was found by Bown, Martin and Potter[23] in an oscillographic study of the distortion in broadcast reception within a given area, and oscillographic observations by Friis[24] of reception on 18.75 megacycles proved consistent with the belief that fading is mainly caused by wave interference.

Very good evidence of real direction shifts was found by Friis[24] using a cathode-ray oscillograph to observe the phase difference of signal waves as received by antenna systems located a fraction of a wavelength apart.

Changes in angle of incidence were found by Friis[24] to be much more frequent and of greater magnitude than direction shifts of the received signals. Heising, by his observations of signal retardation, had previously computed changes in angle of incidence sufficient to indicate rapid variations in the height of the ionized layer, sometimes as great as 72 km (45 miles) in two minutes.

Rotation of the plane of polarization of the indirect ray was suggested from theoretical considerations of the earth's magnetic field and varying ionization in the upper atmosphere by Appleton and Barnett[25] and by Nichols and Schelleng,[26] who indicated that this phenomenon might be expected in transmissions largely along the line of a magnetic meridian. Evidences of abnormal polarization causing apparent direction shifts were pointed out by Eckersley,[27] and Pickard[28] found by many observations made on high frequencies with an analyzer that the plane of polarization under certain conditions rotated 90 deg. in transit. Alexanderson's investigation,[29] also on the high frequencies, seemed to indicate a continuous rotation

[22] P. O. Pedersen, "The propagation of radio waves along the surface of the earth and in the atmosphere," p. 230. Copenhagen, 1927.

[23] R. Bown, De L. K. Martin, and R. K. Potter, "Some studies in radio broadcast transmission," Proc. I. R. E., **14**, 57; February, 1926.

[24] H. T. Friis, "Oscillographic observations on the direction of propagation and fading of short waves," Proc. I. R. E., **16**, 658; May, 1928.

[25] E. V. Appleton and M. A. F. Barnett, "Wireless wave propagation," *Electrician*; **94**, 398, April 3, 1925.

[26] H. W. Nichols and J. C. Schelleng, "The propagation of radio waves over the earth," *Nature*, **115**, 334; March 7, 1925. Also *Bell Sys. Tech. Jour.*, **IV**, 215–234; April, 1925.

[27] T. L. Eckersley, "The effect of the Heaviside layer on the apparent direction of electromagnetic waves," *Radio Review*, **2**, 60–65, and 231–248; February and May, 1921.

[28] G. W. Pickard, "The polarization of radio waves," Proc. I. R. E., **14**, 205–212; April, 1926.

[29] E. F. W. Alexanderson, "Polarization of radio waves," *Jour.* A. I. E. E., **45**, 636–640; July, 1926.

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of the plane of polarization with alternate points of plane and circular polarization along the transmission path. Hollingworth[30] recently made measurements to determine the state of polarization of the indirect ray which produces apparent progressive direction shifts during the sunset period, and on the basis of several assumptions calculated that the plane of polarization of an indirect ray received from a selected transmission of 21 kc (14350 m) rotated 90 deg. during sunset and remained thus abnormally polarized during the night. Eckersley[7] working with the high frequencies, made simultaneous fading records with horizontal and vertical antennas, and reports that under some conditions the fading changes in the two records were so nearly opposite in phase as to indicate rotation of the plane of polarization of the indirect ray.

[30] J. Hollingworth, "The polarisation of radio waves," *Proc.* Royal Society, A, **119**, 444–464, 1928.

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Proceedings of the Institute of Radio Engineers
Volume 17, Number 6 *June, 1929*

Discussion on DETECTION CHARACTERISTICS OF THREE-ELEMENT VACUUM TUBES*

By
F. E. Terman and T. M. Googin

J. C. Warner:[1] The detection characteristics given in this interesting paper have been derived mainly from the static characteristics of the various tubes. This method is accurate when applied to the ordinary type of three-element tube, but in the alkali vapor tube (UX200A) the effect of the positive ion movements, which greatly influence the detector action, is not indicated by the static curves. The mechanism of rectification which takes place in the alkali vapor tube depends upon the amplitude and frequency of the impressed signal voltage. If the frequency is low, for example, 60 cycles, the rectification can be predicted from an examination of the static characteristics. But if a small radio-frequency voltage is applied, of the order of 50 millivolts, the behavior of the tube is quite different. Rectification takes place in such a way as to decrease the plate current even when no grid leak and condenser are used. The static mutual characteristic would, of course, indicate just the opposite. Also, the grid leak and condenser have little effect on the magnitude of detection, and the chief function of the grid leak seems to be in fixing a proper bias.

As the signal voltage is increased the positive ion effects reach saturation and the action then becomes much the same as in the ordinary tube. The conclusions of the paper are therefore justifiable for relatively strong signals but do not apply to the region where the UX200A shows its greatest relative sensitivity in comparison with other tubes, i.e., on weak signals.

F. E. Terman:[2] Mr. Warner is correct in his inference that the results of our paper depend upon static curves. The question with gas tubes is then as to the extent to which the detection in these tubes departs from that given by the static characteristic. I have made some rough measurements at 50 kc which indicate that on an unmodulated voltage at least 50 per cent to 75 per cent of the detector output can be determined by the static characteristic. Just what would happen at 1000 kc I do not know. It will be remembered that the gas tube is used on no commercial sets and, in fact, the Radiola battery sets have always used the 201A tube as a detector in preference to the 200A. In view of these facts I feel that our statement in regard to the 200A tube was not quantitatively justified by the data we had at hand, but that qualitatively it is not very much in error. The unimportance of grid-leak resistance in detection with the 200A is not significant as our static curves indicate that over the entire range of grid leak resistance ordinarily used the detection is equally sensitive.

* Proc. I. R. E., **17**, 149; January, 1929.
[1] Research Laboratory, General Electric Co., Schenectady, N. Y.
[2] Stanford University, California.

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MONTHLY LIST OF REFERENCES TO CURRENT RADIO LITERATURE

THIS is a monthly list of references prepared by the Bureau of Standards and is intended to cover the more important papers of interest to professional radio engineers which have recently appeared in periodicals, books, etc. The number at the left of each reference classifies the reference by subject, in accordance with the scheme presented in "A Decimal Classification of Radio Subjects—An Extension of the Dewey System," Bureau of Standards Circular No. 138, a copy of which may be obtained for 10 cents from the Superintendent of Documents, Government Printing Office, Washington, D. C. The various articles listed below are not obtainable from the Government. The periodicals can be secured from their publishers and can be consulted at large public libraries.

R100. Radio Principles

R113 Störmer, C. Sur un echo d'ondes electromagnetiques courtes arrivant plusieurs seconds apres le signal emis et son explication d'apres la theorie des Auroras Borealis. (On an echo of short electromagnetic waves arriving several seconds after the main signal and explanation by means of the theory of the Aurora Borealis). *L'Onde Electrique*, **7**, pp. 531-33; Dec., 1928.

(Same work reported in abstract in *Comptes Rendus*, **187**, p. 811, Nov. 5, 1921, and in *Nature* (London), Nov. 3, 1921).

R113 Van der Pol, B. Les echos des ondes courtes et les Auroras Borealis. (Short wave echoes and the Auroras Borealis). *L'Onde Electrique*, **7**, pp. 534-37; Dec., 1928.

(Does not agree with Störmer's explanation since rays will never pass through the upper region of the Heaviside layer, but where short waves pass into the layer at a large angle they may go into electron densities where the phase velocity becomes almost infinity, that is, the group velocity almost zero. The reflected wave can then come back to earth after a considerable time has elapsed).

R113 Jelstrup, H. S. J. Essai d'explication de l'echo Störmer-Hals sur les ondes de 31.4 metres de PCJJ. (Analysis of the explanation of the echo of Störmer-Hals on waves of 31.4 meters from PCJJ). *L'Onde Electrique*, **7**, pp. 538-40; Dec., 1928.

(Attributes the long interval echoes to many reflections between the Heaviside layer and the earth and calculates the time lapse between the original signal and the echo by means of Howe's equivalent transmission line formula for the Heaviside layer).

R113 Clapp, J. K. Some experiments in short distance short wave radio transmission. Proc. I. R. E., **17**, pp. 479-93; March, 1929.

(Some experiments in short-wave radio transmission over a distance of 55 miles are described).

R113 Fassbender, H. and Kurlbaum, G. Abhängigkeit der Reichweite sehr kurzer Wellen von der Höhe des Senders über der Erde. (De-

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pendence of the range of very short waves on the height of the transmitter above the earth). *Zeits. für Hochfrequenztechnik*, **33**, pp. 52–55; Feb., 1929.

(Experiments on range of transmission on wavelength of 3.7 m. This work was carried out on an airplane, the transmitter being on the plane and the receiver on the ground.)

R113 Gerth, F. and Scheppman, W. Untersuchungen über die Ausbreitungsvorgänge ultra-kurzer Wellen. (Experiments on the radiation of ultra-short waves). *Zeits. für Hochfrequenztechnik*, **33**, pp. 23–27; January, 1929.

(An investigation of radiation for waves shorter than 28 megacycles. It is shown that the waves spread out like light and that objects larger than a wavelength produce shadow effects. It is assumed that only direct waves play a part and calculations are given for the distance range).

R113.4 Guyot. Étude sur la propagation des ondes courtes. (Study of the propagation of short waves). *L'Onde Electrique*, **7**, pp. 509–530; Dec., 1928.

(Review of the propagation of short waves assuming that the Heaviside layer is defined by the position of the earth's surface with respect to the sun. Based on this, the height of the layer depends on the time during the 24 hours, the latitude of the station and the declination of the sun. Experimental curves are given based on simplified formulas for the calculation of the height of the layer).

R113.5 Maris, H. B. and Hulburt, E. O. Wireless telegraphy and magnetic storms. Proc. I. R. E., **17**, pp. 494–500; March, 1929.

(A recent theory of auroras and magnetic storms attributes these phenomena to the action of a flash of ultraviolet light from the sun; this flash causing an unusual ionization in the Kennelly-Heaviside layer. This theory is found to be borne out in a detailed discussion of data of high-frequency (7500 to 20000 kc) circuits of the U. S. Navy during the magnetic storms of May 28, July 7, Oct. 18 and Oct. 24, 1928).

R125.1 Smith-Rose, R. L. Radio direction finding by transmission and reception (with particular reference to its application to marine navigation). Proc. I. R. E., **17**, pp. 425–478; March, 1929.

(A resumé of the performance of apparatus employed for radio direction finding determination either by transmission or by reception is given. Investigations made in England on this subject and results obtained are reviewed. Application of direction finding to navigation and possible effect of coastal and night errors in connection therewith are discussed).

R125.6 Pistolkors, A. A. The radiation resistance of beam antennas. Proc. I. R. E., **17**, pp. 562–79; March, 1929.

(A method proposed by Brillouin for the calculation of radiation resistance applied to several types of beam antennas is described.)

R130 Kingdon, K. H. and Mott-Smith, H. M. The operation of radio receiving tube filaments on alternating currents. *General Electric Rev.*, **32**, pp. 139–148; March, 1929.

(It is shown that three disturbances may occur in amplifiers using a-c feed for the filament supply, namely, the primary potential ripple, the primary magnetic ripple and the primary temperature ripple. In this part of the paper (part I) the case of the amplifiers is treated and methods given for reducing these effects.)

R132 Brain, B. C. Output characteristics of thermionic amplifiers. *Experimental Wireless and W. Engr.* (London, **6**, pp. 119–127; March, 1929.

(Analytical treatment of the output amplifier for optimum undistorted energy transfer. It is shown that the maximum output is obtained by making the load resistance about 1.6 times the a-c resistance of the tube.)

R133 Lazaref, W. Über die Instabilität der Frequenz von Röhren generatoren und deren Stabilisierung. (On the instability of the frequency of electron tube generators and their stabilization). *Zeits. für Hochfrequenztechnik*, **33**, pp. 55–63; Feb., 1929.

(The effect of filament excitation, "B" voltage and feedback on the frequency is studied. The cause is due to the grid current which increases the decrement).

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R133 Hollmann, H. E. Zusammenfassender Bericht: Die Erzeugung kürzester elektrischer Wellen mit Elektronenrohren. (Compilation on the production of ultra short waves). *Zeits. für Hochfrequenztechnik*, **33**, pp. 27–30, January; pp. 66–74, Feb., 1929.

(Lists the methods for the production of ultra short waves).

R134 Nelson, J. R. Notes on grid circuit detection. PROC. I. R. E., **17**, pp. 551–561; March, 1929.

(Shows that the value of the second derivative of grid current with respect to grid voltage the main term in grid rectification, may be found by the change in direct current in the grid circuit with any desired value of input voltage. An experimentally determined curve is given showing the detector frequency distortion of a commercial set for a 2-megohm grid leak and for a ½-megohm grid leak.)

R134.7 Aigner, F. Das Problem der ökonomischten Vielfachtransponierung. (The problem of economic multi-heterodyning). *Zeits. für Hochfrequenztechnik*, **33**, pp. 9–15, January; pp. 47–52, February, 1929.

(Methods are investigated by means of which the incoming frequency can be changed with minimum number of local oscillators. Two heterodyne oscillators suffice to produce almost any multiple frequency changes.)

R149 Kryter, R. J. Alternating current rectification as applied to radio—Part I. *QST*, **13**, pp. 33–37; Apr., 1929.

(Discussion of tube and chemical rectifiers.)

R200. Radio Measurements and Standardization

R230 Harris, S. An extension of the method of measuring inductances and capacities. PROC. I. R. E., **17**, pp. 516–20; March, 1929.

(Substitution method commonly employed for measuring small capacities is here shown to be a special case of a more general principle.)

R300. Radio Apparatus and Equipment

R334 Pike, O. W. and Spitzer, E. E. A new low power screen grid transmitting tube—UX 865. *QST*, **13**, pp. 43–45; April, 1929.

(Uses of tube as amplifier and characteristics of tube given.)

R343 Beers, G. L. and Carlson, W. L. Recent developments in superheterodyne receivers. PROC. I. R. E., **17**, pp. 501–515; March, 1929.

(Major electrical elements of a modern superheterodyne receiver discussed briefly, and practical volume control is described.)

R357 Freese, H. Beseitigung der Nebenfrequenzen beim statischen Frequenzwandler. (Elimination of side frequencies for the static frequency changer). *Zeits. für Hochfrequenztechnik*, **33**, pp. 1–8, January; pp. 41–46, Feb., 1929.

(The absorption system due to Lorenz (British patent No. 263825) and that due to Zenneck is investigated. As analyser, the cathode-ray oscillograph or an undamped wave meter was used.)

R376.3 Kyle, C. The Kyle condenser reproducer. *Radio Engineering*, **9**, pp. 26–29; March, 1929.

(Practical and theoretical discussion of an electrostatic speaker of interesting design.)

R376.3 Hector, L. G. and Kozanowski, H. N. Apparent equality of loudspeaker output at various frequencies. PROC. I. R. E., **17**, pp. 521–535; March, 1929.

(Describes a type of alternation phonometer which permits rapid switching of power at two frequencies to the same loud speaker without the distracting effects of the transients that would result from ordinary types of commutation. This result is obtained by the use of rotating condensers to provide variable capacitive reactance in the input circuit of the power amplifier that operates the speaker.)

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R381 Coursey, P. R. On the capacity of dry electrolytic condensers. *Experimental Wireless and Wireless Engr.* (London), **6**, pp. 128–132; March, 1929.

(Gives a description of the wet and semi-wet (so called dry) electrolytic condensers and emphasizes that especially for the latter type in alternating current method (not vallistic method) has to be used for measuring the effective capacity.)

R386 Reed, F. Superheterodyne band pass filters. Radio Listeners' Guide and Call Book, pp. 45–47, Spring, 1929.

(Design data for band pass filters.)

R400. Radio Communication Systems

R412 Hull, R. A. Modern practice in high frequency radio telephony. *QST*, **13**, pp. 8–22; April, 1929.

(Discussion of improved method in amateur phone transmission)

R431 Tanner, R. Wm. Reducing noise in broadcast receivers. *Radio Engineering*, **9**, pp. 24–25; March, 1929.

(Use of resonance wave coil, band pass filters and power detector tube.)

R500. Applications of Radio

R520 Marriott, R. H. Electrical aids to navigation (abridgment). *Jour. A. I. E. E.*, **48**, pp. 195–199; March, 1929.

(Electrical and magnetic aids to navigation by water and air are outlined. Discussions on the earth and inductor compass, radio compass, radio beacon, submarine signals, height indicators for aircraft, etc. are given. A bibliography is included.)

R526.1 Dellinger, J. H. Directional radio. *Aeronautical World*, **2**, pp. 20–22; Feb., 1929.

(Description of Bureau of Standards work on radio beacon).

R582 Zworykin, V. Facsimile picture transmission. Proc. I. R. E., **17**, pp. 536–550; March, 1929.

(A facsimile picture transmitting system is described whose chief object is to produce a simple rugged apparatus for practical usage without requiring the attention of a skilled operator).

R800. Non-Radio Subjects

534 Does a vibrating diaphragm carry a mass of air with it? (editorial). *Exp. Wireless and W. Engr.* (London), **6**, pp. 117–118; March, 1929.

(Since dealing mostly with divergent waves the velocity of air particles lags behind the pressure. Hence the force at the surface of a spherical surface will be out of phase with its velocity just as though the diaphragm had a certain mass or as if a mass of air were attached to it.)

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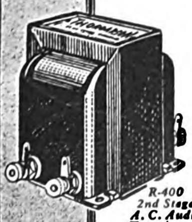

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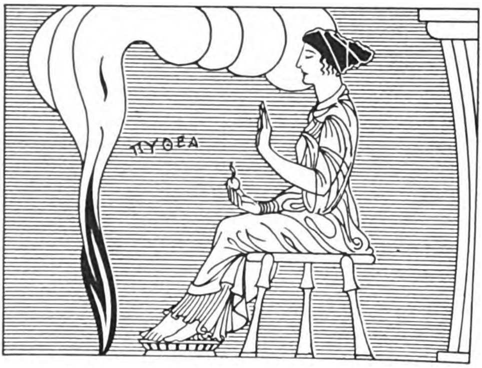

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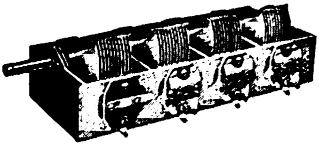

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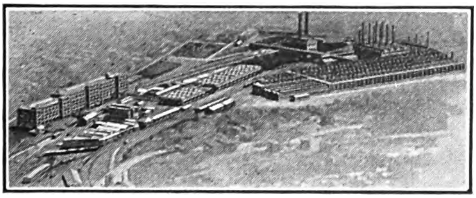

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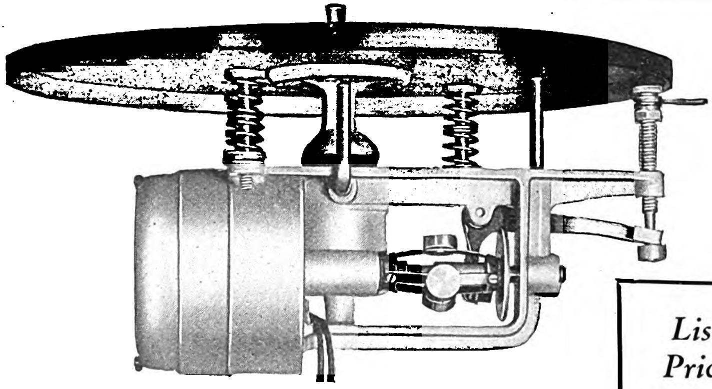

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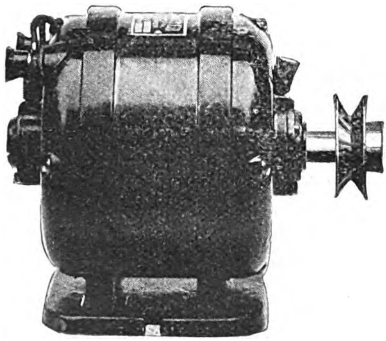

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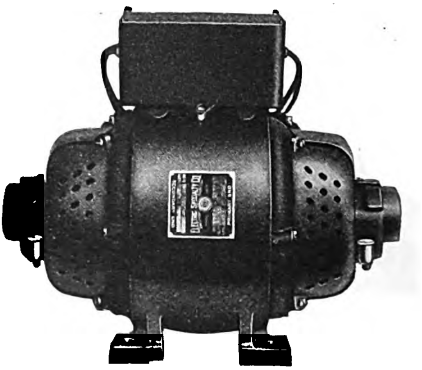

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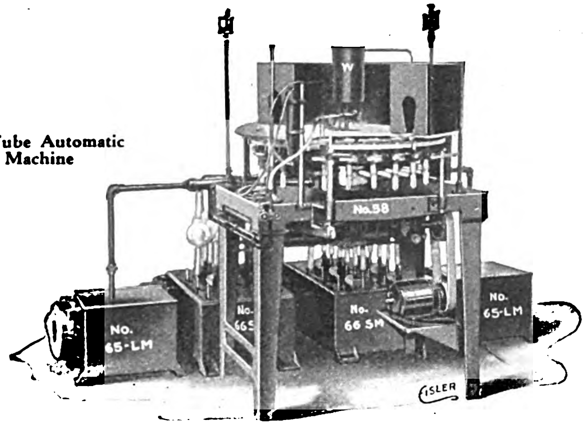

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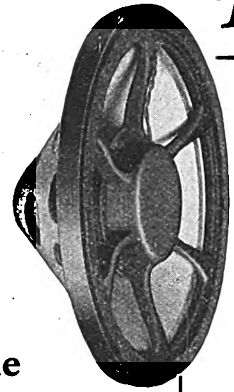

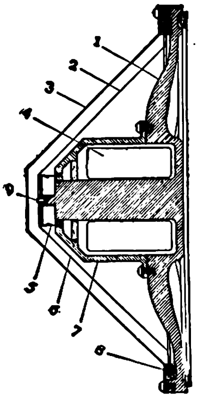

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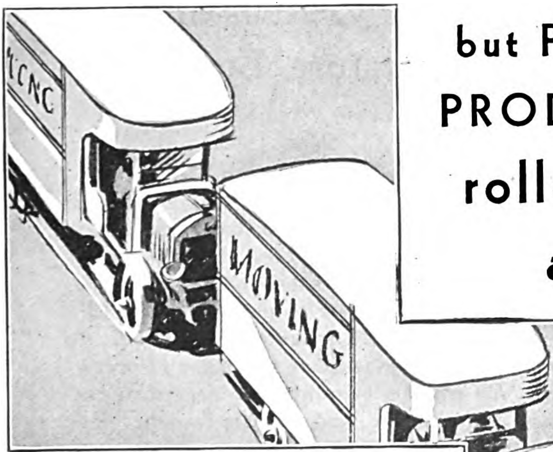

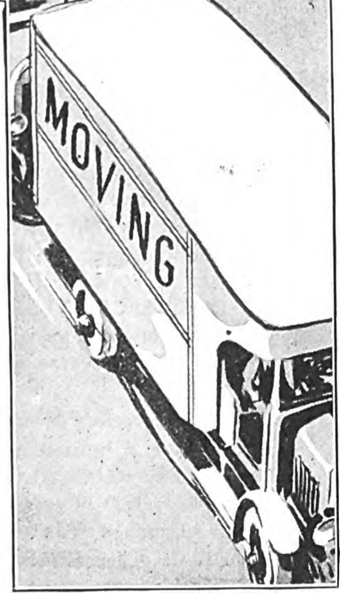

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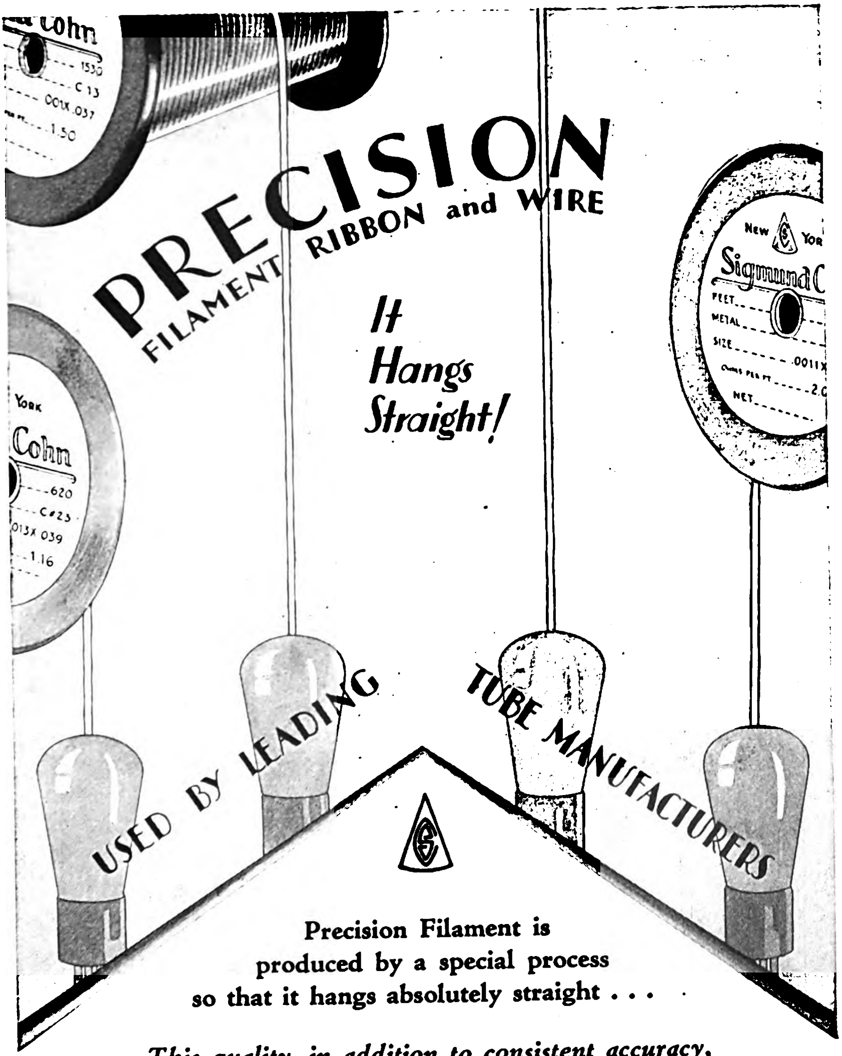

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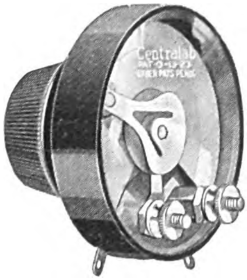
and *Double Junior*



Write for full particulars of specific application.

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CENTRAL RADIO LABORATORIES

36 Keefe Ave. Milwaukee, Wis.

A CENTRALAB VOLUME CONTROL IMPROVES THE RADIO SET

OIL IMPREGNATED FILTER CONDENSERS

REALIZATION THAT THE OIL IMPREGNATED FILTER CONDENSER GUARANTEES FREEDOM FROM TROUBLE HAS INCREASED PROFITS FOR MANY.

SEND US YOUR SPECIFICATIONS WE WILL BE GLAD TO PROVE IT TO YOU.

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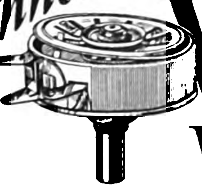
Announcing the CLAROSTAT Wire-wound VOLUME CONTROL



AFTER some two years spent on the serious problem of volume control, the Clarostat Engineering Staff is pleased to present a unique and perfected device.

The CLAROSTAT WIRE-WOUND VOLUME CONTROL combines the high resistance and compactness usually demanded, with maximum reliability and long life. It is a genuine wire-wound resistor—not to be confused with devices utilizing carbonized paper or other uncertain resistive materials that become noisy in operation. An unique form of pressure contact, instead of sliding contact, does away with wear and tear on the finest wire, yet establishes a silent, positive, and velvety-adjustable contact.

Here is the solution of the volume control problem. Here you will find the necessary high resistance; constant resistance at any setting; smooth adjustment; noiseless operation; life-long service if desired; and total freedom from troublesome and costly servicing. The CLAROSTAT WIRE-WOUND VOLUME CONTROL is the hallmark of good engineering in any assembly.

Available in any resistance range. Also supplied in three-terminal or potentiometer style, two-terminal or rheostat style, and in tandem style or duo-volume control for two resistances with a common knob.

Write for complete engineering data regarding the new CLAROSTAT WIRE-WOUND VOLUME CONTROL. If you are a radio set design engineer or manufacturer, write on your firm letterhead for a sample of stated resistance. And remember—there's a CLAROSTAT for every resistance requirement—fixed, variable and automatic.

Clarostat Manufacturing Company, *Inc.*

MEMBER RMA

Specialists in Radio Aids

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BUILT BETTER
CONDENSERS AND RESISTORS

How Do You Buy Condensers?

MOST filter condensers, condenser blocks and bypass units are bought merely on the basis of price, voltage ratings and their ability to withstand ordinary short-time tests, without sufficient consideration to the fact that these are not dependable indicators of the ability of a condenser to stand up under all conditions of service, during the entire life of the receiver or power unit.

Nothing is apt to prove as costly as a cheaply made, over-rated condenser or resistor. Whether you are a manufacturer, professional set builder or experimenter, you cannot afford the high cost luxury of a cheap condenser or resistor.

Aerovox condensers and resistors are conservatively rated and thoroughly tested. The Aerovox Wireless Corporation makes no secret of the Insulation Specifications of their filter condensers and filter condenser blocks. This information is contained in detail in the 1928-29 catalog.

The next time you are in the market for filter condensers or filter condenser blocks, make your comparison on the basis of Insulation Specifications. Aerovox condensers are not the most expensive, nor the cheapest, but they are the best that can be had at any price.

Send For Complete Catalog

Complete specifications of all Aerovox units, including insulation specifications of condensers, carrying capacities of resistors and all physical dimensions and list prices are contained in a fully illustrated, 20-page catalog which will be sent free of charge on request.

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Let us answer the question

"Can you make it of Vulcanized Fibre?"

If you are unacquainted with the physical properties, the machining qualities and the cost of vulcanized fibre applicable to *your* products, put it up to us to find out.

Our experimental department has shown many a busy manufacturer how the proper application of the right grade of vulcanized fibre will overcome serious production or sales difficulties.

Avail yourself of this service now . . . blue prints or a detailed description of the part and its functions will put us to work at once.

Write today!

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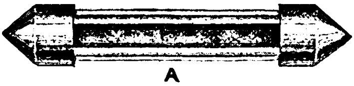
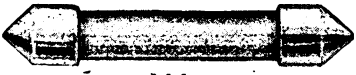
Continental Resistors

Durable dependable, simple in structure and give a minimum of resistor trouble.



A

Type A for grid leaks and light power purposes. Will dissipate ½ watt safely.



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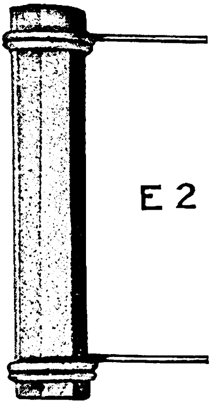
Types W and X for greater power dissipation.

X

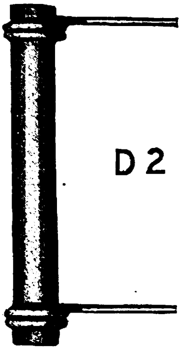
All types furnished in any resistance value desired.

In use continuously for a number of years by the largest manufacturers.

Types E2 and D2 furnished with wire leads soldered to coppered ends, are for soldering permanently into position in apparatus where they are to be used.



E 2



D 2

Samples for test sent on receipt of specifications.

CONTINENTAL CARBON INC.

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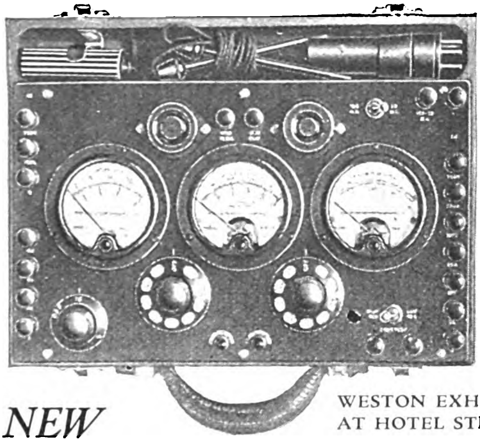

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The NEW

RADIO SET TESTER

See it at the R. M. A. Convention

THE radio industry is familiar with the Weston Model 537 Radio Set Tester—for A. C. and D. C. receivers. Service men hailed it with great acclaim a year ago, noting its many advantages over the Weston Model 519—for D. C. only.

And NOW—here is another great advance—the Weston Model 547—incorporating many additional features to meet the service testing requirements of radio's latest developments. And there have been many since the last R. M. A. Convention. But with this NEW SET TESTER radio servicing is still further simplified, even taking into account the number of new tubes, sets and circuits. Space won't permit description here—nor would words alone do this new set tester justice. You must see it for yourself—operate it—try to think up some service problem it can't solve. Try as you will the Model 547 will give you a quick and accurate answer every time. Convenient—complete—light and rugged. Handsome in appearance—and it will yield you handsome profits. It will increase your business and your prestige. YOU CAN BANK ON IT!

This instrument has many outstanding service features. But first of all it is a Weston—assuring you exquisite workmanship and complete service reliability. It is provided with three instruments—all 3¼" diameter and furnished with bakelite cases. Carrying case, removable cover, panel and fittings are also made of sturdy bakelite.

WESTON ELECTRICAL INSTRUMENT CORPORATION
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Weston
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Performance insured

Radio performance can be no better than the performance of each component part.

The T-C-A standard of quality is your best insurance that these important units, at least, will function as you would have them.

Look us up at the Radio Show and let's talk it over. Booth No. 64D.
Demonstrations at Room 516,
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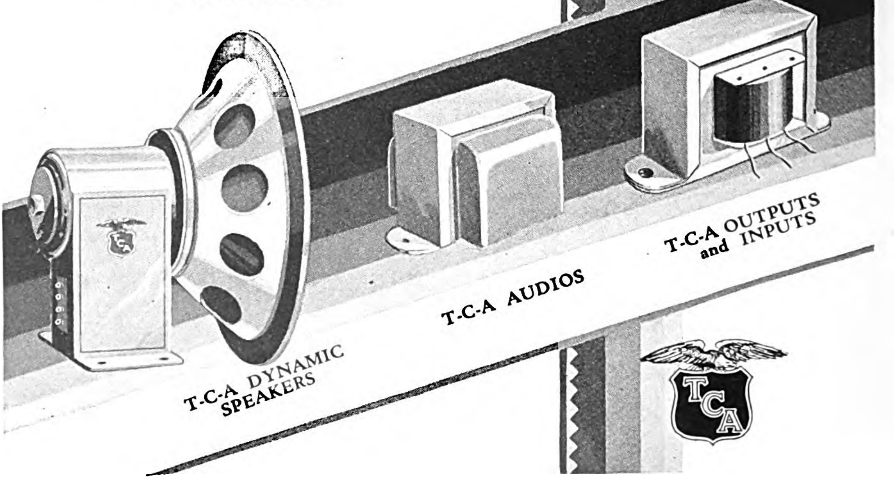

TRANSFORMER CORPORATION OF AMERICA

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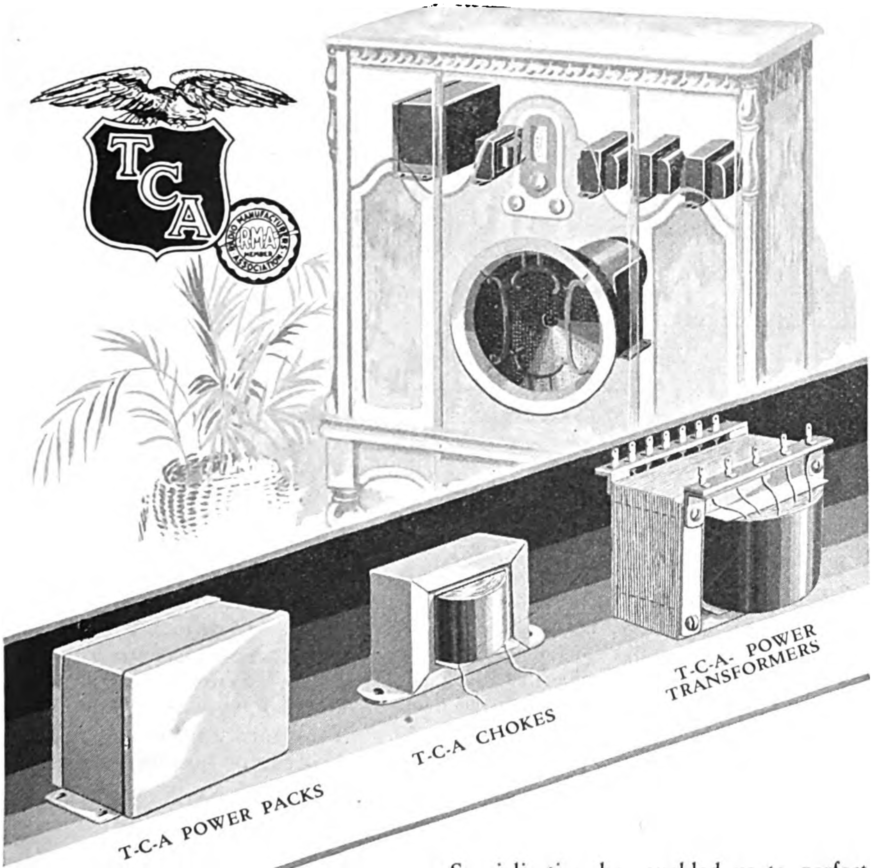

Specialization has enabled us to perfect our product to a point where a large demand has developed. This volume has encouraged us to build and install special machinery, more accurate and more speedy than human hands.

T-C-A Transformers meet the quality requirements of your engineers, as well as the price requirements of your production department.

And the same precision through controlled quantity production that made T-C-A transformers and power packs standard in the country's finest sets, is securing a quality in T-C-A Dynamics that is receiving quick recognition. They are a real contribution to the industry.

Complete data and samples available. T-C-A engineers will gladly assist in your audio and power supply developments.

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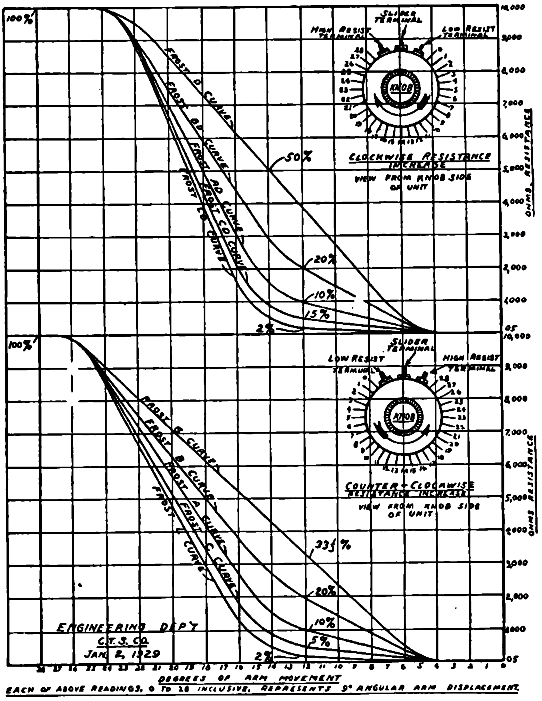
FROST-RADIO Volume Controls supplied in any of these curves, as well as special curves to meet your exact requirements



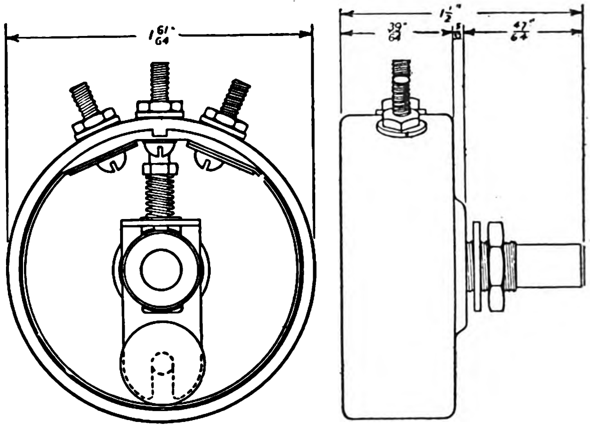
YOU can secure Frost-Radio Volume Controls in any of the curves drawn on the charts shown, as well as in many other curves to suit your needs. You can thus obtain from us any resistance gradient you desire in units from several hundred ohms up to and including a total resistance of several megohms. Frost-Radio Variable Resistors are made in several sizes, in single or tandem construction, and with resistance arranged to increase with either clockwise or counter-clockwise knob rotation. These units are smooth-running, non-inductive, and absolutely unaffected by temperature or humidity changes.

Tell Us of Your Special Requirements

We supply Frost-Radio Volume Controls in the following standard housing dimensions:

Diameter	*Depth*	*Type*
1 61/64 in.	39/64 in.	Bakelite
1 7/8 in.	43/64 in.	Metal
1 7/8 in.	3/4 in.	Metal
1 5/8 in.	37/64 in.	Bakelite
1 1/2 in.	43/64 in.	Metal

For other than standard units, please indicate in your request for samples any special dimensions, terminal positions, curve desired, and watt load, and in requesting tandem unit samples state maximum permissible mounting depth. Please note that bushings, threads per inch, shaft lengths and diameters illustrated are standard. Made in two or three terminal type.



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The Largest Manufacturers in the World of High Grade Variable Resistors

***Main Offices and Factory:* ELKHART, INDIANA**

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JAMAICA PLAIN, BOSTON EST. 1907

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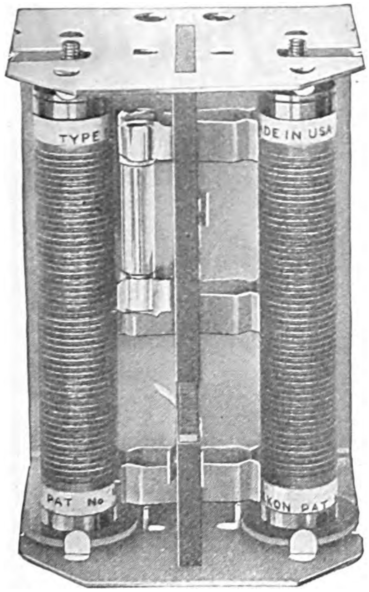
This New Elkon Rectifier Eliminates the Power Transformer in Dynamic Speakers

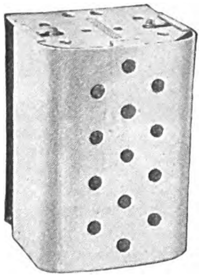


AGAIN Elkon leads the field. The new Elkon D-30 Power Supply is the outstanding development of the year in rectifiers for dynamic speakers. This remarkable rectifier operates directly from the AC power line eliminating the Power Transformer and reducing the cost of assembly.

Supplied complete, ready to install, or the rectifier units (two required on each speaker) can be sold separately.

Wonderfully efficient, quiet in operation. The units can be replaced when necessary as easily as a tube is changed in a socket.

If you have not already sent us a sample of your new speaker, do so at once. We will equip it with the new Elkon rectifier and return it to you promptly.



ELKON, Inc.

Division of P. R. Mallory & Co., Inc.

3029 E. Washington St.
Indianapolis, Ind.

COUPON

ELKON, Inc., Radio Dept. E-72
3029 E. Washington St., Indianapolis, Ind.

Please send us complete information on your new ELKON D-30 Power Supply for Dynamic Speakers.

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Address ....

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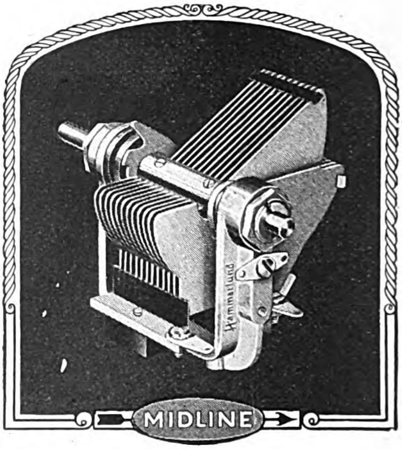
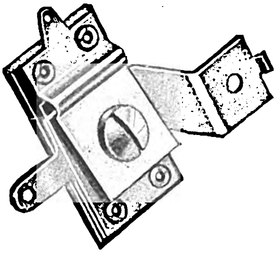
Standard or Special

HAMMARLUND WORKMANSHIP Assures Condenser Perfection





When the standard Hammarlund Equalizing and Neutralizing Condensers do not fit into your receiver design, we are prepared to make special models, either single or in gang, to your specifications.

THE most advanced radio engineering produced the Hammarlund "Midline" Condenser. It has enjoyed more than three years of leadership because it was *born* a leader—the first embodiment of laboratory precision in a stock condenser model.

Sturdy aluminum alloy frame; soldered brass plates, carefully aligned and fixed by tie-bars; cone and ball bearings; phosphor-bronze pig-tail; removable rotor shaft. Individual and gang models. Also the new "Battleship Midline"—the master of multiple tuning units, with sections matched to within ¼ of one per cent.

All desirable gang sizes and capacity ratings.

Write us your needs. Hammarlund cooperation and facilities are yours for the asking. Address Dept. PE6.

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424-438 WEST 33rd STREET, NEW YORK, N. Y.

For Better Radio
Hammarlund
PRECISION
PRODUCTS

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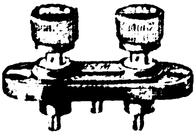
These EBY Products Are Already Half Way On the Set

Their Advanced Designs Reduce Assembly Costs

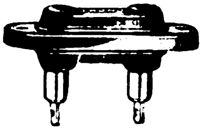
The Eby Company considers assembly cost as part of the cost of the product to the manufacturer. That's why Eby has specialized in designing products that take only half the time and expense to assemble—that's why Eby products are virtually half way on the set when they are finished at our factory.

Another big feature of Eby products is attractive appearance. Every Eby product is an asset to the finest radio set. And finally, and of equal importance, is Eby service. The Eby Company has been an absolutely dependable supplier of quality products for six years.

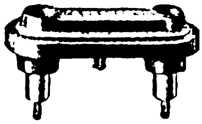
Model 6 Socket is new and *much better*. Designed for manufacturers' use exclusively. Popular Eby features have been retained and new ones added.
Five different colors numbered—for identification of tubes.
Long two-sided prongs—for positive contact.
Guide for tube prongs—a famous Eby feature.
Rivet assembly—for economy
New in performance, new in appearance—and a new low price



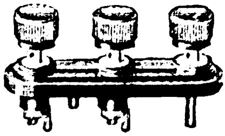
Here's another assembly shortcut for manufacturers who want binding posts for Speaker Field connections. One nut to tighten — two soldered connections to make and they are both assembled and insulated. No washers!



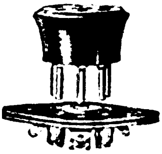
A pair of tip jacks moulded as inserts, in a brown Bakelite strip marked Speaker, Phono or Field. No insulating washers or nuts! Eby type H



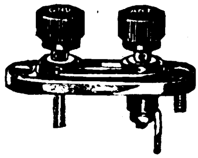
One nut assembles these two tip jacks moulded as inserts in a brown Bakelite strip. Available marked Speaker, Phono or Field. No insulating washers! Type "S"



Another product which eliminates assembly operations. The Ground post is automatically grounded and the two Antenna posts insulated with no washers. Quick, easy, economical!



A new and inexpensive connector plug and receptacle. Available with six prongs for connecting set to power supply, with five prongs for push-pull dynamic speaker voice coil and field connections or with four prongs for dynamic speaker voice coil and field connections



There are three assembly operations on the Combination Antenna and Ground post strips — two nuts to tighten and one soldered connection to make. The ground post is automatically grounded to panel and Antenna post insulated from it. No washers! Furnished with soldering lug and nut assembled on Antenna post

The H. H. EBY MFG. CO., Inc.
4710 Stenton Ave. Philadelphia
Makers of Eby Binding Posts

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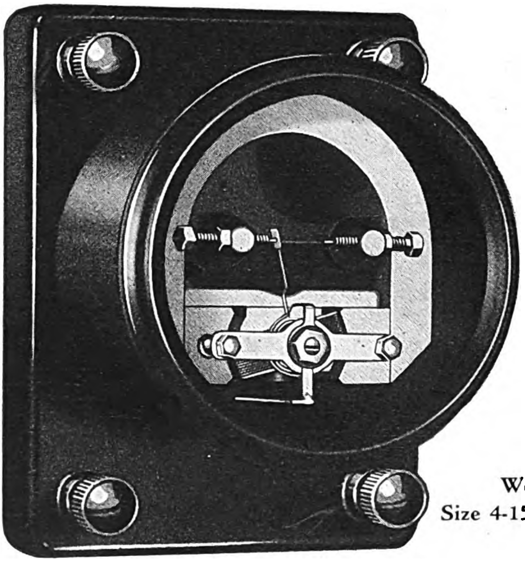

Jewell

Pattern No. 142

Relay

Weight 3 lbs.
Size 4-15/16" x 6" x 4¼"

Making accurate tests . . . with automatic equipment

AUTOMATIC equipment makes possible the utmost accuracy in high speed testing of complicated electrical apparatus, such as radio sets.

In automatic testing a sensitive relay functions on a very small change in current or voltage, completely eliminating the inaccuracies of the human element.

Automatic equipment is now being used extensively in testing vacuum tubes, condensers, chokes, and transformers.

The Jewell Pattern No. 142 Relay is particularly suitable for automatic testing. A special pole piece construction in coordination with a husky cobalt steel magnet gives a high gap density. This, in turn, results in a torque many times that obtained from a commercial instrument, even for the small values of plate or grid current encountered in the testing of radio equipment.

Pattern 142 Relays can be supplied to function on a current as small as 50 microamperes, and with high and low contacts which are adjustable in service within certain limits.

All applications of automatic testing are more or less special, and Jewell Engineers invite correspondence regarding your problems.

JEWELL ELECTRICAL INSTRUMENT CO., 1650 Walnut St., Chicago

Manufacturers of a complete line of high grade switchboard and portable instruments

29 YEARS MAKING GOOD INSTRUMENTS

JEWELL

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Potter Condensers

Quality - Long Life - Economy

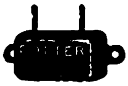
Potter Filter Blocks

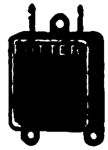
Built to meet present day requirements ,insuring long life and hum-free eliminators. Each unit in these condenser blocks is made of the highest grade materials and must meet the high standard set by Potter. They are carefully and permanently assembled and hermetically sealed in metal containers.

All are very conservatively rated, as to voltage breakdown, insuring a trouble-free condenser so important for complete consumer satisfaction. Condenser Blocks can be assembled in different capacity combinations to meet requirements.

Every item entering into the manufacture of Potter Condensers is the very best that can be secured, assuring a product that will exceed your highest expectations.

Potter By-Pass Condensers

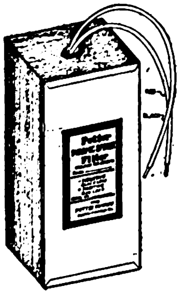




Designed for by-pass work and their high quality and conservative rating permit their application to circuits where high "B" voltages are employed giving satisfactory results.

They are hermetically sealed in metal containers—with the best sealing compound obtainable, creating a condenser that is moisture proof with uniform capacity.

The condenser mounting lugs have uniformly placed holes for bolting or riveting to a die-punched panel assuring a rigid and permanent assembly.



Potter Dynamic Speaker Filter

Will reduce the hum to a minimum when used with the A.C. operated Dynamic Speaker using the low voltage rectifier.

The construction allows the use in any position as it is of the dry type which is a solid mass in texture, permanently sealed, and in no way should be confused with electrolytic wet condensers using solutions, pastes or jellies, and requires no servicing attention. The dielectric provides the durability and minimum leakage current, which, in a unit of this kind, is of the order of one milliampere at ten volts in the correct sense according to the polarization.

A Potter Dynamic Speaker Filter is the ideal unit to use with a dynamic speaker to correct whatever hum may be present and to give increased operating qualities.

The Potter Co.

North Chicago, Illinois

A National Organization at Your Service

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Type HTD RADIO CONTINUITY TESTER

The new RADIO CONTINUITY TESTER was designed, and is now offered, as an ideal instrument for making a wide variety of tests on radio receiving sets, in particular, and on many other radio devices and circuits.

One of the largest makers of radio receiving sets in the Country has found that the Continuity Tester will, without any other instruments, perform the majority of the essential tests on its many types of sets.

Send for Supplement No. 1 to Bulletin No. K-300

"Over thirty years' experience is back of Roller-Smith"

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Electrical Measuring and Protective Apparatus

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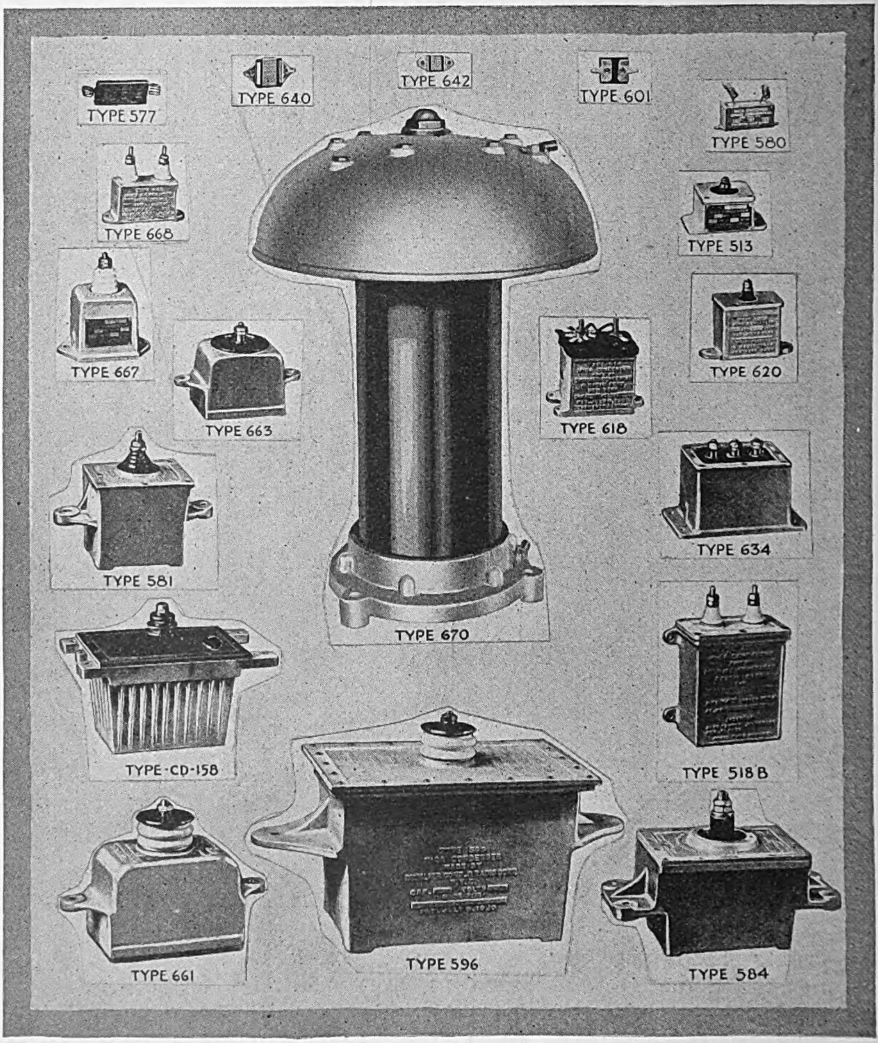
You can forget the Condensers if they are Dubilier's



JUST a few of the thousands of types of condensers developed by Dubilier Engineers for radio and industrial purposes. Ask for catalog 89.

Dubilier

CONDENSER CORPORATION

342 Madison Avenue

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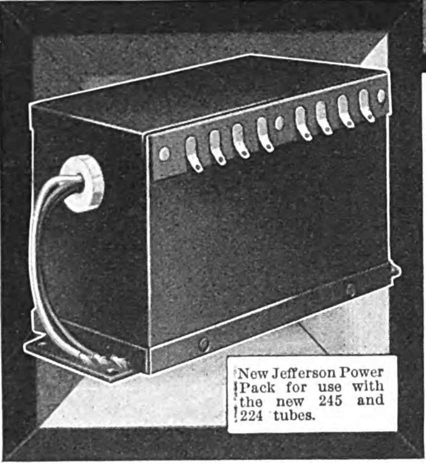
Engineering Co-operation *to* SOLVE *your* TRANSFORMER PROBLEMS



New Jefferson Power Pack for use with the new 245 and 224 tubes.

KEEPING step with the progress of electrical development, Jefferson has maintained a reputation for quality transformers . . . for engineering co-operation in designing and developing transformers for special application.

With the advent of radio, a large and complete engineering department, a research laboratory and a staff of sales engineers were added to render definite assistance in the solution of electrical problems.

Today, numerous radio manufacturers attribute much of the success of their sets, from an electrical standpoint, to the help of Jefferson engineers in the design of their audio and power transformers. Likewise, they have benefited by Jefferson production capacity—which insures prompt deliveries during peak seasons.

These are the services which Jefferson offers you, too—in addition to serving as a reliable source of supply for quality transformers. Our engineering and research departments are maintained to serve you. Let us know your problems.

JEFFERSON ELECTRIC COMPANY

1591 S. Laflin Street Chicago, Ill.

Transformers and Chokes for New Power Tubes

As specific evidence of Jefferson engineering progressiveness, we present the new power transformers, designed for use with the new 245 power tube and the 224 shield grid tube. To work with these new transformers, we have a wide range of choke units—heavy single duty chokes—double choke units of conventional design— or staggered choke units, one heavy and one light choke, an especially economical method which minimizes hum and allows maximum voltage on power tubes without overloading the rectifier. Special audio transformers, improved in design, are also available to make use of all the possibilities of these new tubes.

* * *

Complete electrical specifications and quotations on these new units will be furnished on request.

JEFFERSON

AUDIO and POWER TRANSFORMERS and CHOKES

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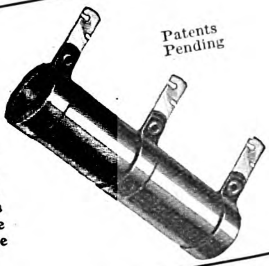
For the manufacturer who is proud of his product!

New Five-watt Volume Control with Longer Life

U. S. Pats. No. 1034103 —1034104 and Pats. Pending.

Study this remarkable resistance mechanism—the resistance element fused to an enameled metal base—floating silver contact. Smoothness, current carrying capacity and ENDURANCE never before possible. Completely enclosed in metal cover for rapid heat dissipation. Sample, laboratory graphs and full information to established manufacturers.

You'll Praise This New and Better Covered Resistance



Patents Pending

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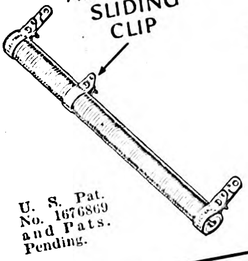

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Manufacturer ....

Address .... City .... State ....

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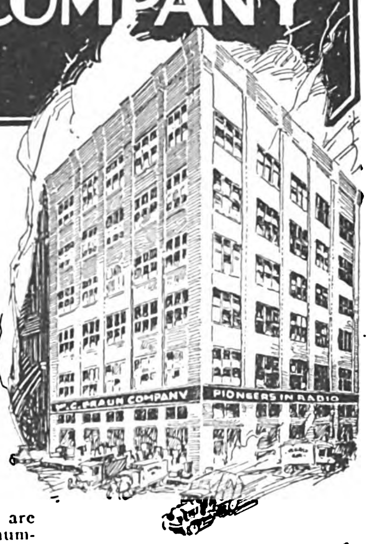

Keep Abreast of the New Developments in Radio

No industry in the world's history has attracted so many inventors and experimenters as the radio industry. Something new is always on tap. Contrast the old wireless days with the modern electrically operated talking radio. Think of what is still to come when perfected television, telephony, short wave control, etc., are fully realized.

In keeping with the policies of *Wholesale Radio Headquarters* (W. C. Braun Company), our service lies in testing out and determining which of these newest marvels are practical, salable and usable for the greatest number. Our task is to study the multitude of new merchandise, select those items that are thoroughly proved and reliable, and make it easy for the public to secure these while they are estill new.

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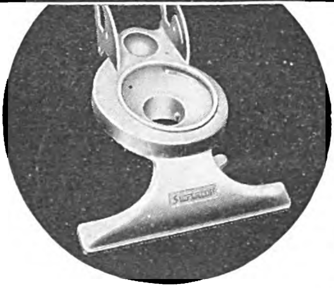

This Alumac Al Die Casting, is used in the Sturtevant Vacuum Cleaner, combines Nozzle, Fan Housing and Regulator. Note deep undercut of fan housing.

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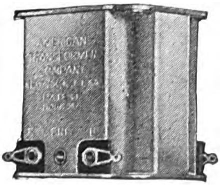

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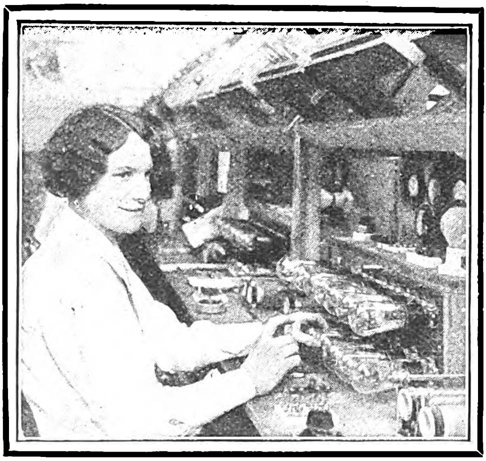

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NOT very long ago a prominent radio manufacturer unrolled a blue-print, and turning to the Dudlo sales engineer who stood by his desk, said: "Here's a job for you fellows at the Dudlo plant. We've had a lot of trouble with this power coil. Can't seem to get it to deliver the proper "B" voltages for these new tubes without overheating. Here's where we need your help . . . what can you do for us?"

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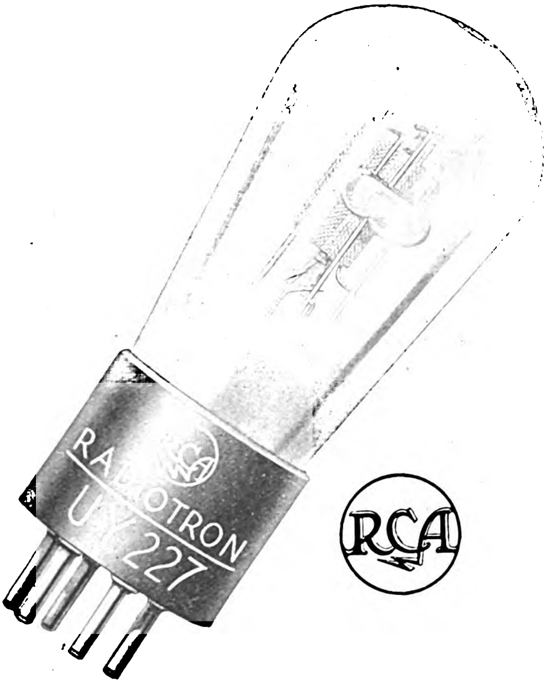

The public looks for the RCA mark on the vacuum tubes of a new radio set. That is one important test of the quality of the instrument. RCA Radiotrons are standard equipment throughout the radio industry.

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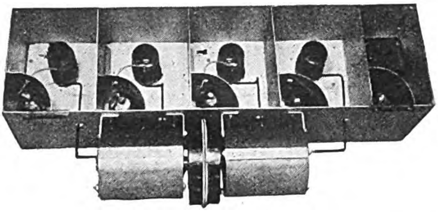
"SEVEN SEAS" CONSOLE

A new design by C. R. Leutz using 3 A/C screen grid radio frequency stages, all tuned; power detector; initial audio stage and push pull power audio amplifier for either two 210's or two 250's. Split, single dial control; 100% scientific shielding; dynamic loud speaker and an adjustable selectivity device.

The result, a night range loud speaker volume of from 500 to 1,000 miles using an indoor aerial only a few feet long, great selectivity and the very finest quality of reproduction.



View of the radio frequency and detector chassis showing the efficient shielding arrangement resulting in a minimum of losses. Condensers also totally shielded. Power audio amplifier and power pack are in separate chassis.

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Cables "Experinfor" N.Y.

LITERATURE ON REQUEST

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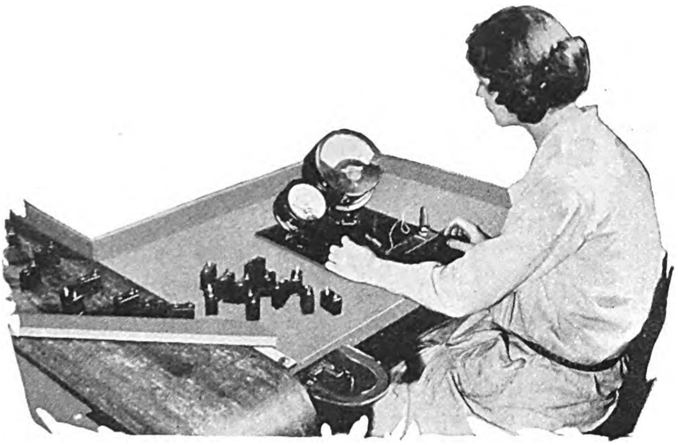
How Your Condensers Are Tested



THIS picture from the FAST condenser factory shows the testing equipment at the end of each conveyor belt system. The completed condensers are progressively conveyed by belt system to the testing table, where severe tests are made by expert operators before the product is ready for shipment.

The painstaking care taken in the various processes of production insure to you a uniform quality product throughout. After the testing operation is completed, the condensers are carefully packed in individually marked cartons, ready for shipment to the radio factory.

This is but one of the many unit operations to which *Fast* condensers are subjected in the process of manufacture.

Put your condenser problems up to the FAST Organization of condenser specialists.

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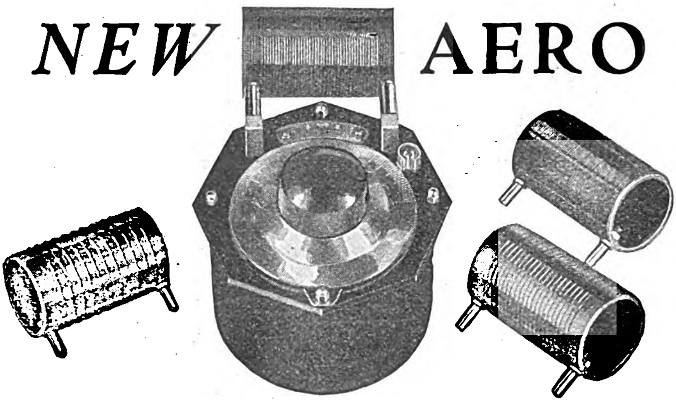

WAVEMETER

THE AERO WAVEMETER was designed for the radio amateur and the experimenter. It is of rugged mechanical and good electrical construction and meets the exacting requirements of the properly conducted amateur station.

Using the principle of the "series gap," and having a definite fixed minimum capacity, the amateur wavelength bands are spread over a great many dial divisions. The velvet vernier dial has 100 divisions, each of which may be read to one-tenth of a division. The 5-meter band covers 40 dial divisions; the 10-meter, 40 divisions; 20-meter, 17 divisions and the 40-meter band, 25 divisions. For the 80-meter band, two coils are used. One (72 to 82 meters) covers 45 divisions, and the other (80 to 90 meters) covers 65 divisions..

Each coil, excepting the 5-meter, uses No. 18 enameled wire tightly wound into grooves of the bakelite tubing. The tubing has a 1/8 in. wall and is 2 in. in diameter. The Aero heavy duty plugs and jacks make positive contact. The heavy brass case has a black crackle finish. The 3/16 in. black bakelite top is hexagonal shape to prevent rolling. The indicator is a standard neon lamp. Each wavemeter is individually hand calibrated. Standard coils are 20, 50 and 80 meters. 5- and 10-meter coils extra. Furnished with or without neon lamp indicator. Price, without neon lamp indicator.. **$24.00**

Aero Listening Monitor

The Aero Listening Monitor enables any amateur transmitter to check his own note, to know whether it is pure D.C. or not. It is a safe, sure and accurate way of knowing your station without depending upon the reports of others. Watching the ammeter won't help you—you need the Aero Monitor to be absolutely positive. The Aero Monitor is a completely shielded unit encased in a black metal cabinet, size 9 x 5½ x 2½ in. Filament and B supply is included. It operates with a UX-199 type of tube and employs a stable circuit, delivering a signal intensity of about R-4 or 5. Equipped with automatic filament control. The battery supply is thoroughly shielded from the R.F. A reliable, scientific instrument. Ship. wt., about 2½ lbs.

Model M-29 Aero Monitor. Including Dry Batteries, but no tube. List Price.$15.00

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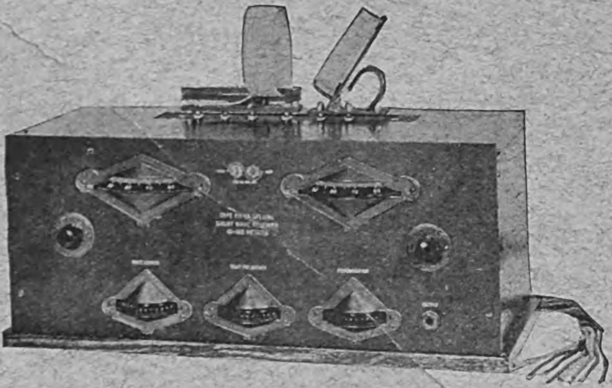

The Grebe CR-18 Special Short Wave Receiver

Short Wave Quality
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WHEN the technically trained man works with any piece of apparatus, precision and accuracy must characterize that equipment. It is a real tribute to the Grebe CR-18 Special that so many nationally-known engineers use it constantly.

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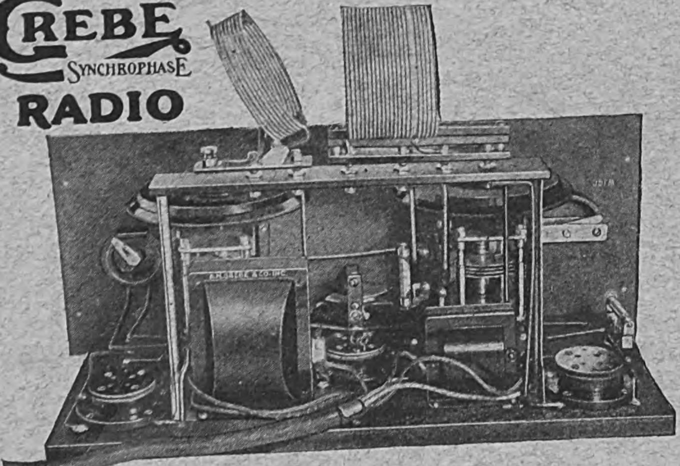

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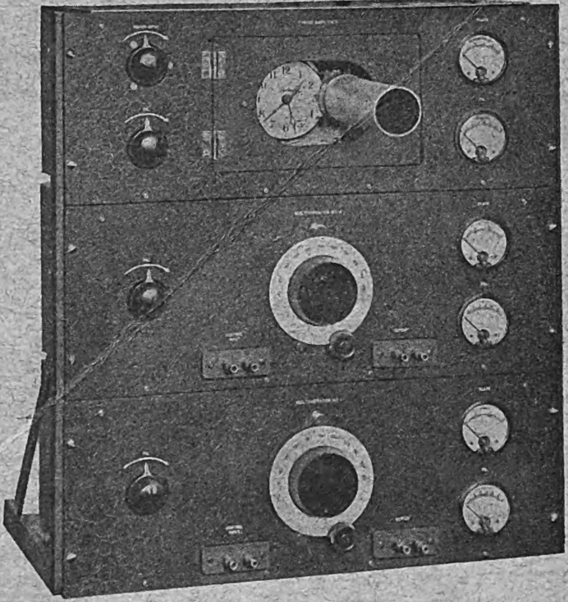
A Frequency Standard



WE have developed, for use in the precision measurement of frequency, a radically new frequency standard. It is, in fact, a complete frequency determination system with the following features:

1. It provides the user with a continuously-operated standard that maintains its frequency with an average deviation of about one part in a million.

2. It provides means for measuring the average frequency of the standard by comparing it directly with the standard time interval defined by radio time signals. The time comparisons can be read to within 1/100th of a second.

3. A large number of frequencies between one kilocycle and 3,000 kilocycles per second are continuously available for making comparisons in the laboratory. Each of these is known with the same precision that the frequency of the standard is known.

4. It is rugged, compact, and reliable, and can be built to meet the requirements of the user.

Please mention the type of service for which the standard would be used when writing for details.

General Radio Company

Manufacturers of Electrical and Radio Laboratory Apparatus

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